WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

JUNE 1954

VOL. 31 No. 6 THREE SHILLINGS AND SIXPENCE

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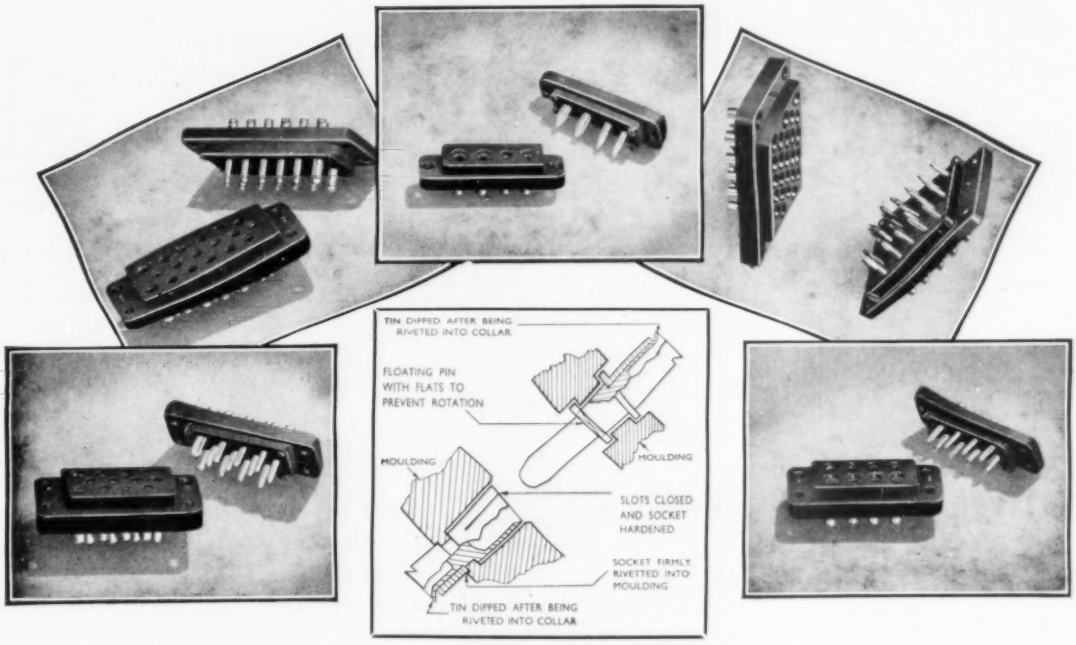
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MOULDING
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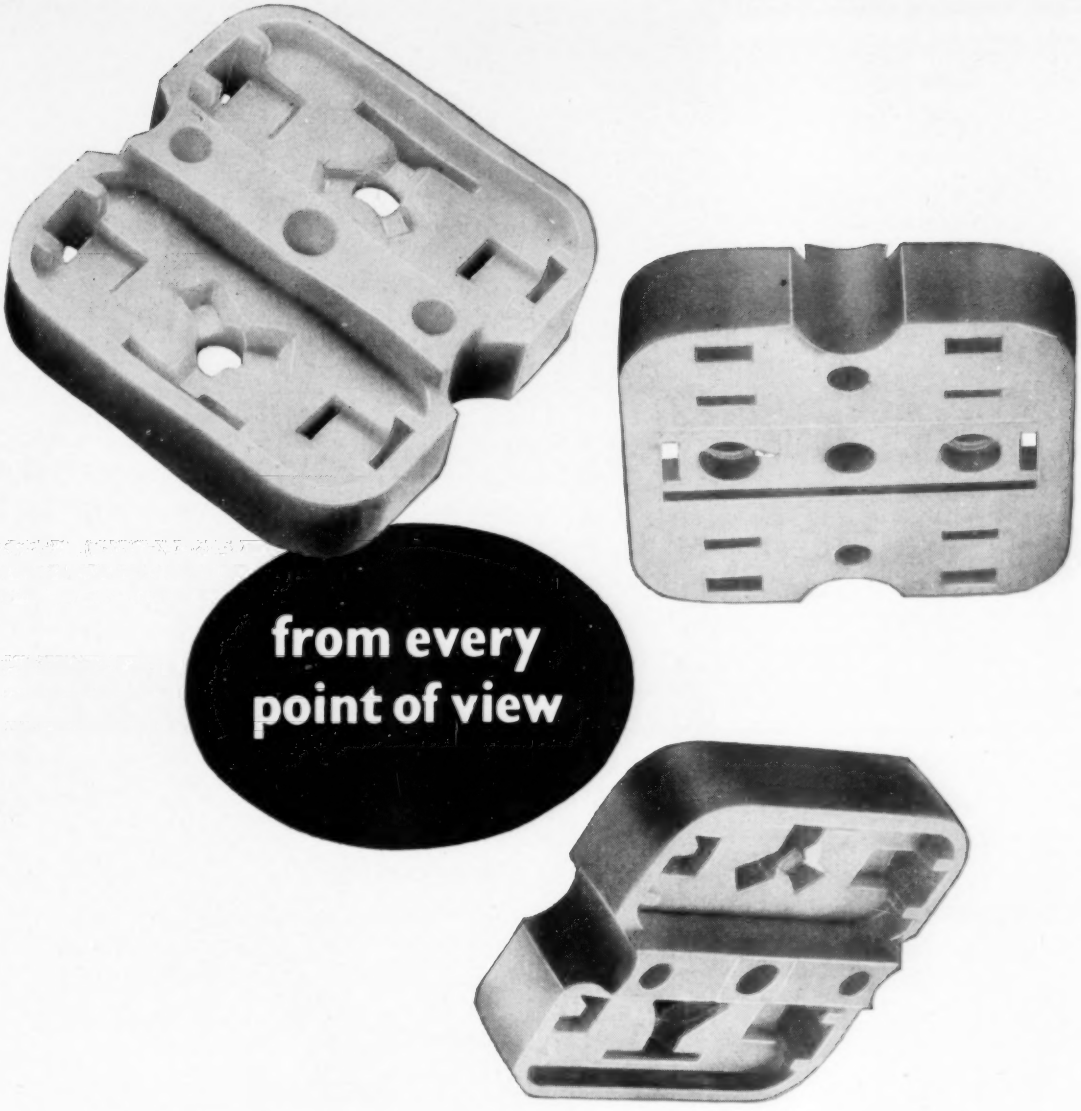

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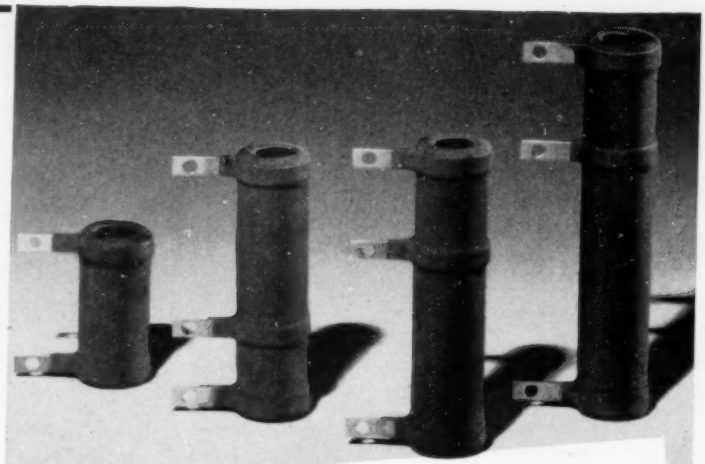
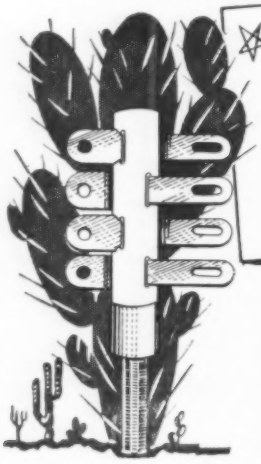


TYPE	31	32	33	34
WATTAGE	10	18	25	40
RESISTANCE OHMS	10–6000	10–12000	10–15000	10–30000
LENGTH	1.9/16″	2.13/16″	3.1/16″	3.13/16″
DIAMETER	11/16″	11/16″	11/16″	11/16″
MAX. TAPS	1	2	3	4



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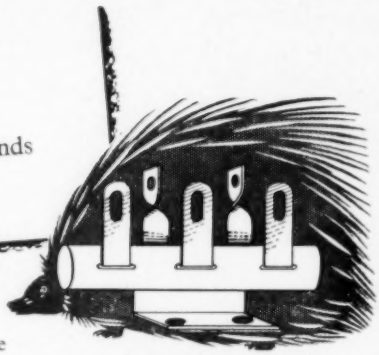

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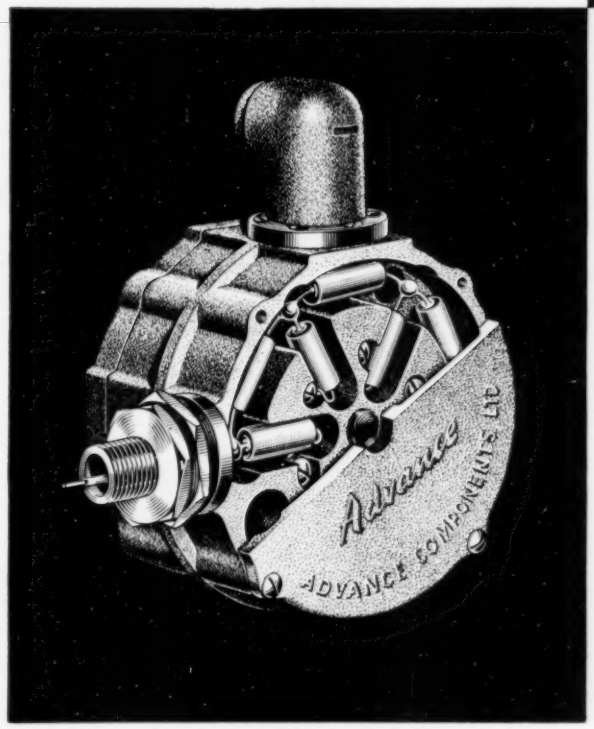

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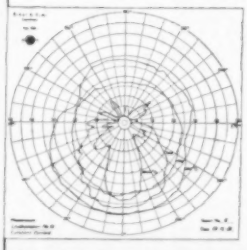

As shown in the photograph the Polar Diagram Recorder is attached to the level recorder section of instrument type 2314. Turntable type 2371 for the test object, connected with a flexible shaft to type 2370, can be easily driven in case of remote operation, by means of synchros or magslips.

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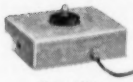
Type 2314 Type 2370



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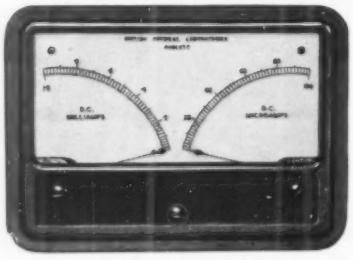

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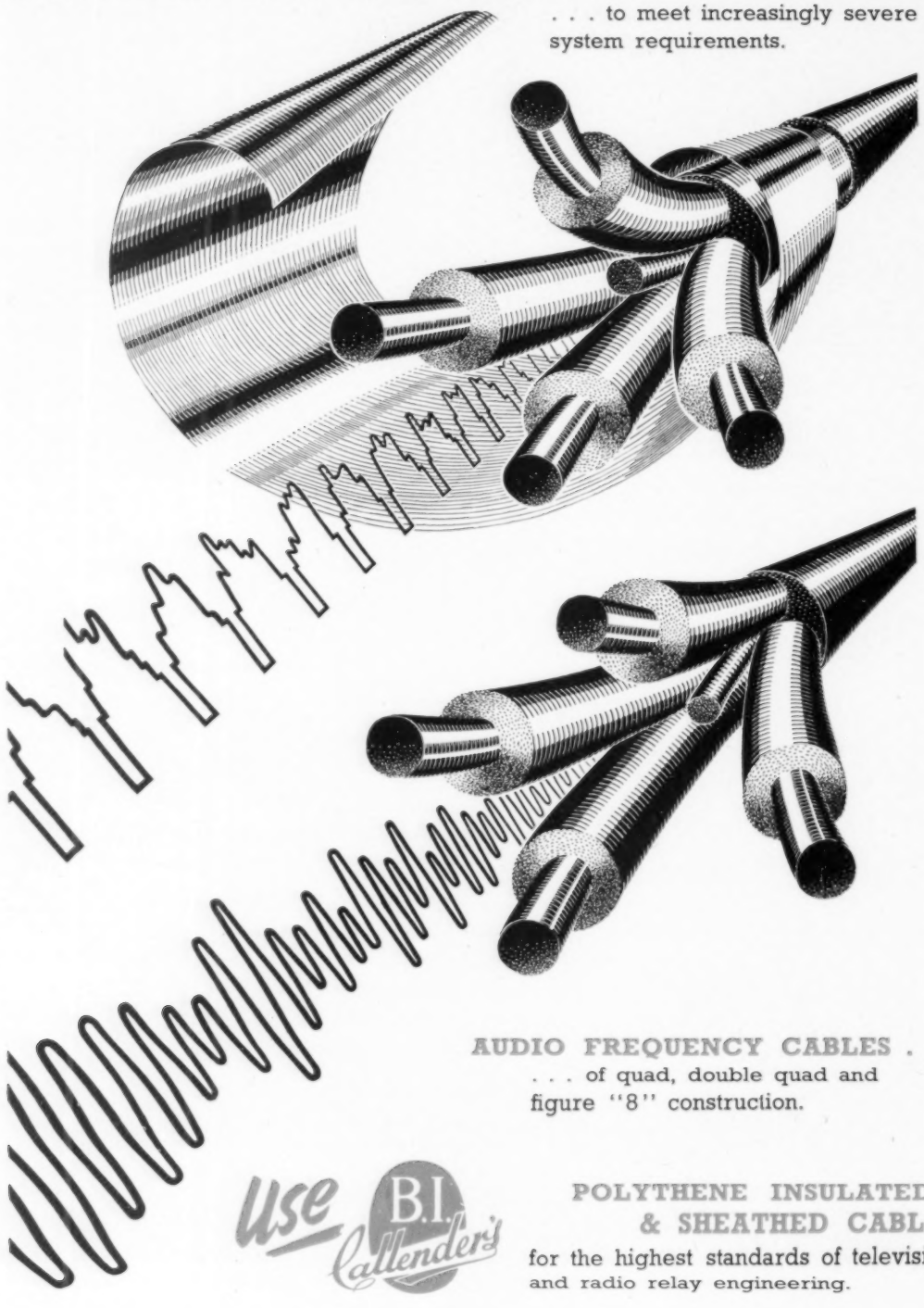

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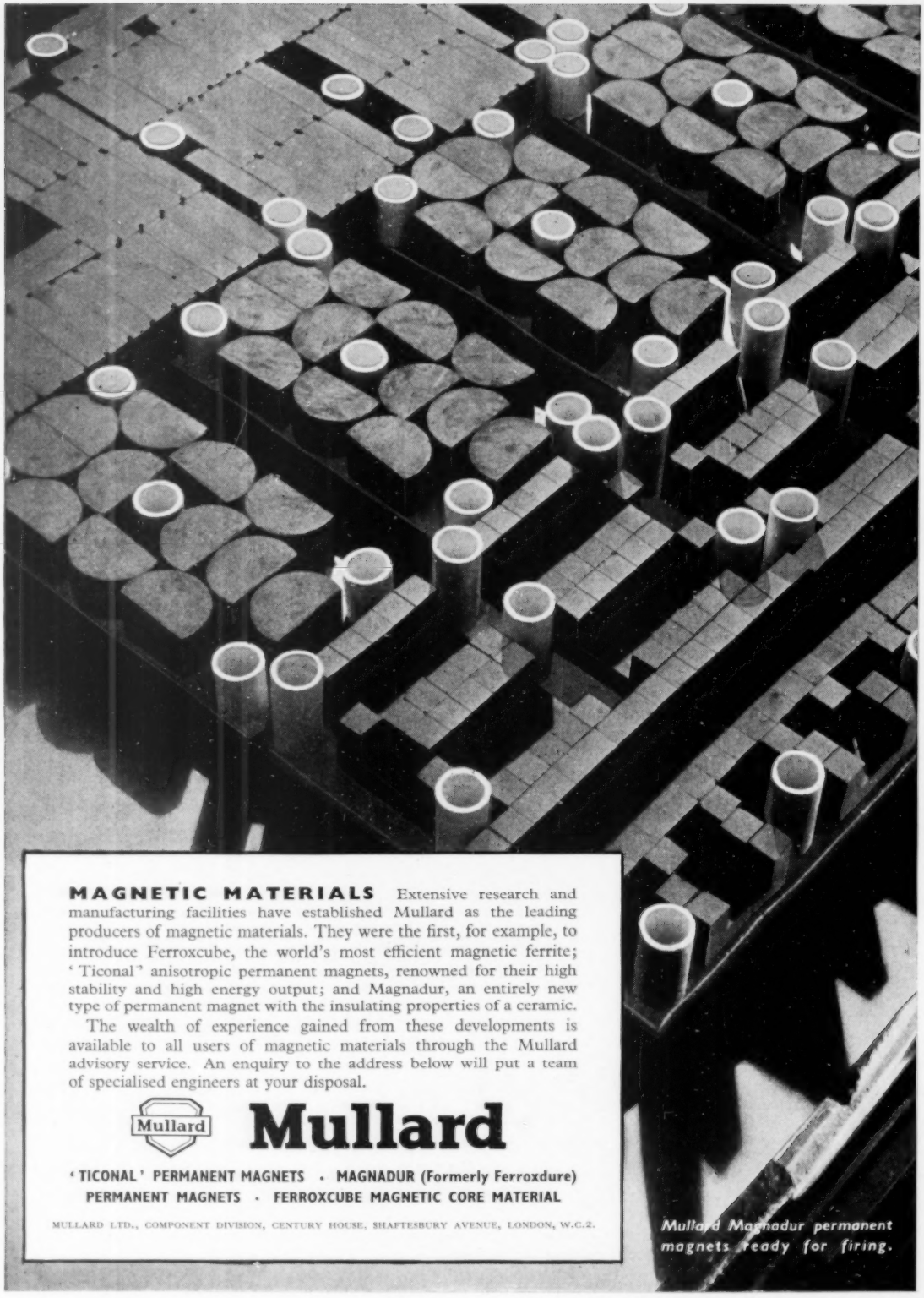

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OA 73

All glass

POINT CONTACT

GERMANIUM DIODE

ABRIDGED DATA

Forward Current:

at +0.25V	....	> 100μA
at +1.0V	....	> 8mA

Reverse Current at 25°C:

at −1.5V	....	< 20μA
at −10V	....	< 200μA
at −20V	....	< 800μA
at −30V	....	< 1.25mA

Operating Ambient Temperature: −50 to +75°C.

R.F. Characteristics: the OA73 is suitable for operation as a wideband R.F. detector.

Fusion sealed

AGAINST MOISTURE

Penetration of water vapour is undoubtedly one of the major factors which affect the stability and reliability of germanium diodes. Partial and complete sealing techniques, while keeping the inside of diodes moisture-free, have in the past often introduced other undesirable fillings.

The Mullard aim has been not only to safe-guard diodes from the ill-effects of water vapour, but also to reduce the number of different materials used in construction and so minimise the risk of contamination. The ultimate point in the evolution of this type of construction is marked by the OA73, the first of a Mullard range of *all-glass* diodes.

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WIRELESS ENGINEER

Vol. 31 JUNE 1954 No. 6

Electromagnetic Induction from Different Points of View

THE divergency of views that are held on this subject was brought out in a paper recently read before the I.E.E. by Mr. P. Hammond* and in the subsequent discussion. The middle section of the paper was entitled "Electromagnetic Induction" and it is to this that our remarks will be confined.

If Fig. 1(a) represents an electron coming towards us in an unchanging magnetic field, it is equivalent to a current element in the opposite direction and will be associated with a right-handed circular magnetic field which will distort the uniform field in such a way as to cause a downward force on the electron, which is therefore deflected. Fig. 1(b) is much the same except that the electron under consideration is one of the free electrons in the conductor which is moving towards us; the electron will experience a downward force which is equivalent to an e.m.f. in the upward direction. In the Faraday disc armature the circumferential movement of the free electrons causes them to experience a radial force, which, in the absence of any external circuit, will cause a potential difference between the centre and the periphery of the disc. Current will therefore flow in any circuit connected between brushes on the axis and periphery. It is amusing to see the efforts often made to explain this fundamentally-simple generator by means of the variation of magnetic flux through fictitious circuits.

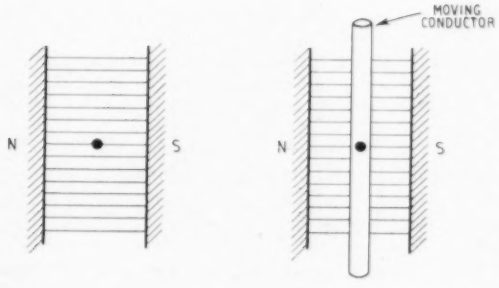

Fig. 1.

Fig 2(a) shows a bar magnet and a conductor moving relatively to each other. To an observer on the magnet the e.m.f. induced in the conductor is due to its movement through an unchanging magnetic field. To the same observer the e.m.f. induced in the rectangle is due to the difference

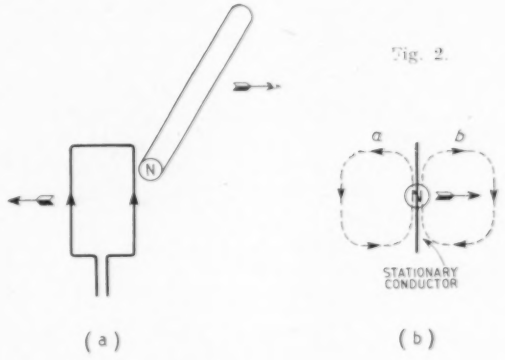

Fig. 2.

* A short modern review of fundamental electromagnetic theory. Read 11th Feb., 1954.

between the e.m.fs induced in the two sides because of the difference between the fields through which they are moving, or he might explain it as being due to the decreasing flux through the circuit. This really comes to the same thing, for as the field is everywhere unchanging, the root cause of the e.m.f. is the movement of the conductors in the magnetic field.

But how would this be explained by an observer on the conductor? To him the conductor is at rest and the magnet moving and causing variations in the strength of the magnetic field in all the surrounding space. If we consider the two closed paths in Fig. 2(b), as the pole moves to the right, the flux coming towards us through the loop *a* decreases, whereas that through loop *b* increases, and the e.m.fs induced in the two paths are as shown. The conductor is thus situated in an electric field and, in the case of the rectangle, one side is in a stronger electric field than the other. To one observer the space around the magnet is a simple stationary magnetic field, whereas to the other observer it is a combination of changing magnetic and electric fields.

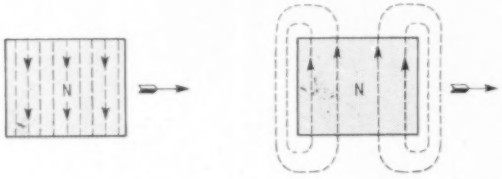

Fig. 3.

One can cover both cases by picturing the lines of force and saying that the induced e.m.f. is due to the relative motion of the conductor and the lines or the cutting of one by the other, but this seems to lack fundamental explanation. To the observer on the magnet nothing is happening inside the magnet, as the field is unchanging, but to the observer on the conductor the magnetic material is moving through its own magnetic field and therefore inducing an e.m.f. within the material as shown in Fig. 3(a), but if one considers two closed paths embracing the leading and trailing edges of the magnet, e.m.fs will be induced in them as shown in Fig. 3(b) due to the changing flux. These two e.m.fs, or the charges that they would produce on the surface, cancel each other, and so give the same result as obtained by the other observer. In all the cases so far considered the e.m.f. is due to the motion of the conductor or of the magnet and can presumably be called motional e.m.f.

Turning from linear motion, which is purely relative, to rotation, which is absolute, let us consider the simple 2-pole dynamo with a smooth armature shown in Fig. 4. Neglecting the armature coil, and assuming the poles to be excited and the armature rotating, the field will be everywhere unchanging, both in the air-gap and in the iron, to the normal stationary observer.

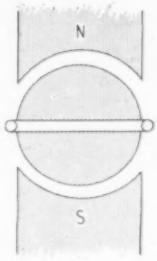

Fig. 4.

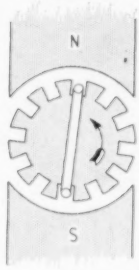

Fig. 5.

If, now, the armature coil is rotated, either alone with the iron armature at rest, or attached to the rotating armature, the coil is moving in a magnetic field which at no point is undergoing any change either in magnitude or direction with respect to time. If, now, the coil is stationary and the polar system is rotated in the opposite direction, so that instead of the coil cutting the lines the lines cut the coil, the result is, of course, the same, and the e.m.f. in both cases can surely be called motional.

In his paper Mr. Hammond deals with this in a very peculiar manner. He divides the field of the stationary poles and the armature into two components B_1 and B_2, the former due to the field magnets and the latter to the surface polarity of the armature. He then says "a purely motional e.m.f. will be induced in the armature coils by their rotation through the field B_1. The component B_2 does, however, experience a time variation *which would be noticed by an observer travelling with the armature coils.*" (The italics are ours). He concludes that "the e.m.f. induced in the coil is thus found to consist of a motional effect due to the movement of the coil through the constant field of the pole magnets, and a transformer effect caused by the variation of the surface polarity of the armature". This is surely quite unjustifiable. Although at any point in the rotating armature fixed with respect to the stationary observer, one iron molecule is being continually replaced by another, the magnetic condition at the point is unchanging, and the surface polarity of the armature, as viewed from the pole, is as stationary as that of the pole itself, and neither B_1 nor B_2 undergo

any time variation. His observer travelling with the armature coil—in other words, spinning round with the armature coil—would certainly note a time variation of B_2, but also of B_1. To him they would both, like the poles themselves, appear to be rotating in the opposite direction to that in which he is rotating. The coil is really rotating through a field which to a stationary observer is undergoing no change whatever at any point and the e.m.f. is purely motional.

The conditions are entirely different if the armature is slotted, for the magnetic field at every point in or near the air-gap is changing as the armature rotates, and the induced e.m.f. can be explained in different ways. If one regards the lines of force (or magnetic induction) crossing the air-gap as definite physical entities, sometimes crowded together in a tooth and sometimes widely separated in a slot, then, although a conductor in a slot is in a very weak field, it must cut all the lines emanating from the pole as the slot moves across the pole-face. This it does due to the very high speed at which the lines flash across the slot, a speed many times the peripheral speed of the armature. From this point of view the induced e.m.f. is certainly motional. If, however, we adopt a less materialistic view of the magnetic field, we can explain the induction of the e.m.f. in a different way. The conductors are moving in a very weak field and the motional e.m.f. so induced will be small, but as the armature in Fig. 5 rotates, the flux through the coil, which will be very small in the position shown, will pass through zero and then increase as the teeth on one side of it pass from the N to the S pole and the teeth on the other side from the S to the N pole. The variation of the flux through the coil is thus due to the growth and decay of the flux through the teeth without any flashing across the slots. If the armature were fixed and the poles rotated in the opposite direction, the coil would not be moving, even in a weak field, and the induced e.m.f. would be due entirely to the variation of the magnetic flux through it. Should this be called a motional or a transformer e.m.f.? G. W. O. H.

External Field of a Widely-Wound Toroid

IT has been suggested to us that we might devote an Editorial note to the following question. Fig. 1 shows a toroidal coil wound, say, on a non-magnetic core; the dotted circle is a closed path external to the toroid, and the line integral of H around this path should be zero. If, however, the turns are rather widely spaced, as shown in the figure, and the dotted circle is not much bigger than the toroid, a unit pole being carried round this path would come within the field of the outer conductors as it passed near them, and it is difficult to see how the line integral

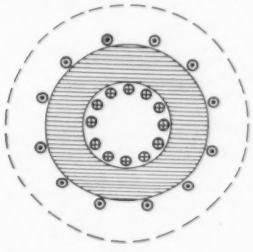

Fig. 1.

could be zero. The answer to the question is seen in Fig. 2, which shows a small piece of the toroid; it is perhaps simpler to assume a cylindrical toroid so that the wires are straight and parallel. Considering the field near the outer surface, Fig. 2(a) shows the field due only to the outer wires, Fig.

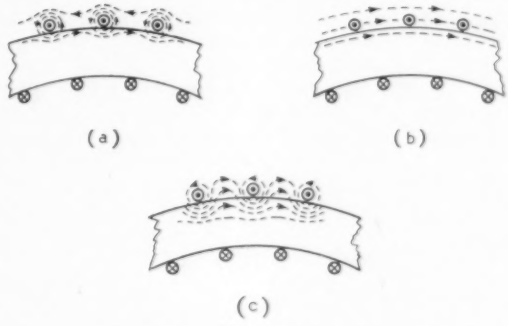

Fig. 2.

2(b) the field due only to the inner wires, and Fig. 2(c) the resultant field. In (a) the field between the wires is very weak, being zero midway between them so that the resultant field there is mainly due to the inner wires. From Fig. 2(c) it is clear that in following the dotted path close to the toroid one passes alternately through fields in opposite directions, thus giving zero line integral. Even if the spacing is not very wide, the same must be true to some extent, because, if, on passing very near the wire, its field can be detected, there must be a reversal of the field in passing from one wire to the next since the line integral must be zero. G. W. O. H.

CURRENT-NOISE IN COMPOSITION RESISTORS

By D. A. Bell, M.A., Ph.D. and K. Y. Chong, M.Sc.

(Electrical Engineering Department, University of Birmingham)

SUMMARY.—Current-noise appears generally to be a resistance-modulation effect, but in composition resistors the expected relation $\overline{\delta I^2} \propto i^2$ is experimentally found to be valid only for small currents. At large currents there are hysteresis effects in the noise without comparable effects in the d.c. resistance. It is shown that the noise is very sensitive to current concentration, and this could account for the lack of correlation between noise and d.c. resistance with large currents.

1. Dependence of Noise on Steady Current

THE view that current-noise in semiconductors is a resistance-modulation effect implies that the mean-square fluctuation voltage δv^2 should be proportional to the square of i, the steady current. Since there are also reasons (both from dimensional arguments and from the type of mechanism proposed by Macfarlane in 1947) for supposing that the noise would be proportional to i^{x+1}/f^x the relation between noise and steady current is of theoretical as well as practical interest.

The first set of measurements on carbon-composition resistors showed the noise power to increase less rapidly than as the square of the current, Fig. 1, but closer investigation of the noise with small steady currents, Fig. 2, showed that the linear relationship $\delta v^2 \propto i^2$ was valid for steady currents producing a dissipation of one-tenth or less of the nominal rating of the resistor. (See also Templeton and MacDonald, 1953.)

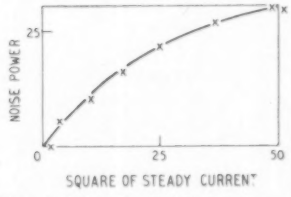

Fig. 1 (above). *Relation between noise and current for large values of current.*

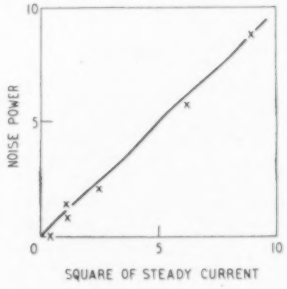

Fig. 2 (right). *With small currents the noise power is nearly proportional to the square of the current.*

Although there are considerable differences in behaviour between different specimens of nominally similar resistor (see below), Fig. 3 is a typical example of a noise characteristic up to a loading of two-thirds of the nominal rating. These characteristics showed differences between successive runs; they could also be changed, often with a substantial reduction of noise, by subjecting the resistors to an overload of six times the nominal dissipation for ten minutes. At the same time the steady voltage versus current characteristic remained constant and ohmic (Fig. 4). From the occurrence of these hysteresis effects it is believed that the departure from the initial proportionality between noise and square of steady current is not due to non-linearity of the essential physical law but to partially-reversible discontinuous changes in the internal structure of the resistor. (An explanation of these changes in terms of thermal instability has been proposed by one of the authors, Chong, 1953).

2. Importance of Current Density

In a homogeneous material it would be better to express the noise in terms of *current density* and *specific resistivity*, rather than in the customary manner in terms of *current* and *resistance*. Composition resistors are not homogeneous bodies, so that (a) the current-density in the conducting paths within the resistor is far greater than the value obtained by averaging over the whole of the cross-section of the body, and (b) the current-density may vary widely between different points along a given conducting path between the two terminals. Owing

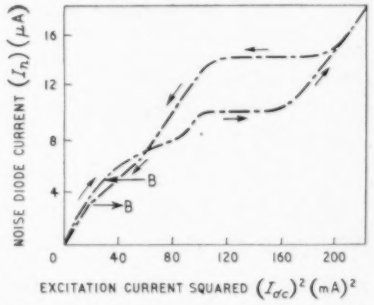

Fig. 3. *Typical example of a noise characteristic.*

MS accepted by the Editor, October 1953

to the dependence of the noise on the *square* of the current-density, the effect of the series combination of the unequally-loaded parts will be to make the noise disproportionately dependent on the high-density parts of the path. If the noise originates mainly at comparatively few points where the current-density is high, any thermal or other modification of the structure at these points can greatly change the noise without much affecting the d.c. resistance, in the way found in the experimental results of Figs. 3 and 4.

This view was strikingly confirmed by an experiment designed to show how a redistribution of current paths can drastically change the noise output from a resistor while having only a small effect on the d.c. resistance. Split circular clamps were attached to the outside of the resistor at the ends, Fig. 5, and the 'noisiness' of different systems of connections was compared by adjusting the steady current in each case to give a reference value of noise output, and the results are shown in Table 1. (The identification of terminals is by means of the letters shown in Fig. 5.) It will be seen that the connection of

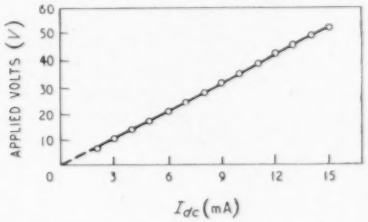

Fig. 4. *Relation between applied volts and current.*

the clamped connections in parallel with the normal terminals makes no significant difference to the resistance but increases the noisiness about 50 times. (The noise is nominally proportional to the square of the d.c., so that reducing the latter seven-fold is equivalent to a 49-fold change of noise in a given resistor.) The further increase of noisiness when clamped terminals alone are used is out of all proportion to the accompanying change of resistance. Since the clamps were making only mechanical contact with the untreated surface of the resistor, it seems likely that the electrical contact was confined to a limited number of points on the surface, thus giving rise locally to high current densities and high noise level.

TABLE 1

Current flow	D.C. for ref. value of noise	Measured resistance (ohms)
A to B	7 mA	3,800
A + C to B + D	1 mA	3,800
C to D	26 μA	5,800
A to B + D	53 μA	3,800
A + C to B	62 μA	3,800

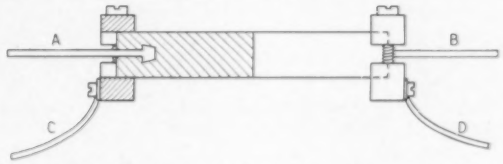

Fig. 5. *A and B are the normal resistor connections; C and D are extra clip contacts.*

3. Variability of Composition Resistors

A difficulty which appears to be common to other types of semi-conductor also (see, for example, the measurements quoted by Torrey and Whitmer on crystal rectifiers) is that different specimens of nominally similar resistors show widely different values of current-noise coefficient. It might be expected that there would be a random variation and, if there were, the dispersion within a group would lead to a straight-line plot on probability graph paper* (taking 'random variation' to mean that the distribution about the mean is Gaussian or follows the so-called 'normal law of errors'). Results for a sample of 20 nominally-similar resistors are plotted in Fig. 6, curve (a), where there appear to be five resistors which are consistent (points within the dotted rectangle), one which is exceptionally good, and 14 which are scattered over rather a wide range of higher values of noise. Earlier results for a sample of six only,

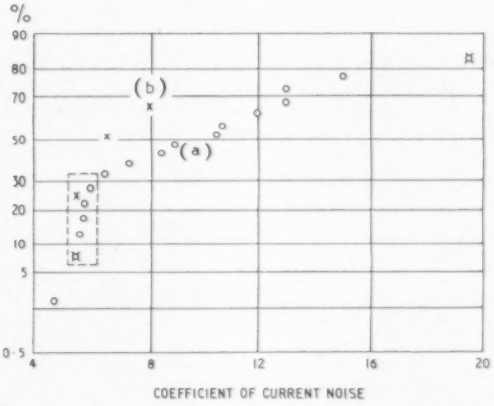

Fig. 6. *Performance of 20 nominally-alike composition resistors, curve (a) and earlier results for six resistors only, curve (b).*

* 'Probability graph paper' as a test for Gaussian distribution was devised by A. F. Dufton (1930) and its use has been reviewed by W. R. Hinton (1949).

curve (b), show a similar trend, and in the graph published by Campbell and Chipman (1949) there is (in the only case for which they record measurements on 20 samples) a tendency for some points to be grouped together near the minimum noise for given conditions, while others spread out to much higher values of noise: this is in their Fig. 4(a), for 1-W metallized resistors of 11,000 Ω. In comparing different types of resistor it is therefore desirable to measure 20 specimens of each, and a suitable representative value to take is the mean value from the closely-spaced group as indicated in our Fig. 6 for group (a). Unfortunately the first set of measurements covered only six specimens of most of the types concerned, and in these instances the lowest individual value of noise has been taken to represent the basic state.

Acceptance of the local-current-density source of noise makes this variation between nominally similar resistors less surprising, for the nominal similarity refers to the overall resistance while the noise, according to the local-density theory, is localized in regions which may be insignificant from the point of view of total resistance.

4. Relation between Noise and Bulk of Composition Resistors

Measurements were made on a range of different types of resistor all nominally of 3·3 kΩ; and the noise coefficient was taken to be the number of microamps of noise-diode current which was equivalent by substitution to the excess noise in the resistor per squared-milliampere of steady current. It was originally supposed that the noise would be inversely proportional to the total volume of the resistor, and that there would probably be a shape factor as well. From a qualitative consideration of the combination of conducting paths in series and in parallel it was thought that for a given value of resistance the noise coefficient might be a function of l^2/Av where l is the length, A the sectional area and v the volume; and it was found empirically that the results for resistors of the same resistance value but different shapes and sizes did fit this formula rather better than several other functions

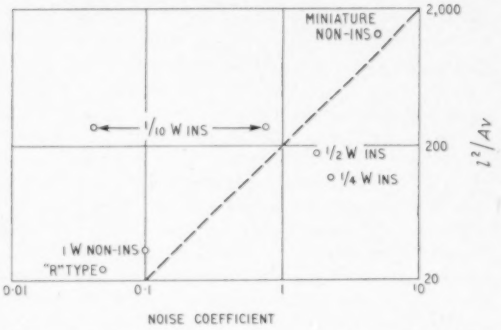

Fig. 7. *Relation between resistor performance and its shape factor. Each point on this diagram is deduced from measurements on a number of samples.*

which were tried. This empirical relationship is illustrated with log-log scales in Fig. 7, where it is clear that there is one type which is very much better than the others and presumably this type has some radical difference in manufacture from the other 1/10th W, which is externally similar but produces as much noise as the average.

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INTERMODULATION DISTORTION IN RECTIFIER MODULATORS

By D. G. Tucker, D.Sc., M.I.E.E., M.Brit.I.R.E.

(Royal Naval Scientific Service)

SUMMARY.—It is shown that non-linear distortion occurs in linear rectifiers, and balanced modulators using linear rectifiers, due to the interference of the signal voltages with the switching function of the carrier. On the assumption of signal voltages low compared with the carrier, a method of calculating the harmonic and intermodulation-distortion products is developed, and a full tabulation is given, for two input signal tones, of all third-order distortion terms based on first- and third-order modulation. An appendix extends this in general form to all orders of modulation and to second- as well as third-order distortion. The values actually tabulated apply to a single rectifier with finite forward and back resistances, and to an ideal ring modulator and ideal balanced shunt and series modulators in which the carrier-generator resistance is zero, and the rectifiers have zero forward and infinite back resistance. For more practical cases, the functions expanded in this paper can be readily fitted into a previous analysis by Belevitch of harmonic distortion in modulator circuits which have finite rectifier and carrier-generator resistances; they thus enable the intermodulation products to be obtained for these circuits. But except in the balanced shunt and series circuits with high carrier-generator resistance, the departure from the performance of ideal modulators is small.

It is shown that the non-linear behaviour described above is not similar to that caused by non-linearity in the rectifier resistances, and cannot be represented by the same equivalent circuits except by an artifice in which different non-linear coefficients are taken in calculating the products of the various orders of modulation, and in which the independent variable of the non-linear equation is the ratio of signal to carrier voltages instead of the signal voltage itself.

LIST OF SYMBOLS

As used by Belevitch:—

F = carrier frequency (c/s)
f = signal frequency (c/s)
v_c = instantaneous carrier voltage
v_s = instantaneous signal voltage

As used by present author in Reference 8:—

p = angular frequency of carrier
$p - q$ = angular frequency of signal
so that q = difference frequency
$E_1 \cos pt$ = carrier voltage
x = ratio of signal/carrier voltage across the rectifier or any circuit containing it which is common to both signal and carrier.
h_1 = coefficient of $\cos pt$ in the Fourier expansion of the modulating or switching function of a single rectifier. This is, in effect, the factor representing the influence of the external circuit on the first-order modulation efficiency. In particular, if $r_f = 0$ and $r_b = \infty$, then $h_1 = 2/\pi$ for any finite value of the external circuit resistance.
r_f = forward resistance of rectifier
r_b = back resistance of rectifier
k = order of modulation; i.e., the order of F in the output components $kF \pm mf_1 \pm nf_2$.

1. Introduction

THIS paper is intended to supplement existing accounts of non-linear distortion in rectifier modulators, and will not attempt to summarize the whole subject. The literature of the subject is not large, however, and is comprised mainly of four contributions in *Wireless Engineer* by Belevitch[1,3,4] and Tucker and Jeynes,[2] together with a few others[5,6,7]. In these papers it is shown that non-linear distortion occurs in modulators such as the ring, shunt and series types (which are balanced rectifier-modulators with a high-level carrier voltage applied longitudinally), due to two main causes:

(a) The non-linearity of the rectifier characteristics when biased either forwards or backwards (i.e., the slope or a.c. resistance is not constant and therefore varies over the cycle of superimposed signal voltage).

(b) The interference with the switching of the rectifiers from forward to backward condition and vice versa—an operation ideally under the sole control of the carrier voltage—caused by the superimposition of the signal voltage.

The first cause is dealt with by Christiansen[6] and is not inherent in the conception of the modulator action. The second cause is inherent in the use of any carrier waveform other than a square-wave; in normal usage, a sinusoidal carrier e.m.f. is provided. Belevitch[1,3,4] has given a very comprehensive and elegant analysis of this effect, but has restricted his consideration to distortion occurring when the applied signal voltage is a single sinusoid.

It is the purpose of the present paper to show how the distortion type (b) can be calculated when the signal comprises two or more tones. As in Belevitch's work, it is assumed that the rectifiers are 'linear', that is to say, they have a constant back resistance r_b and a constant forward resistance r_f, and switch from one to the other at zero applied voltage.* The basis of the

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* The problem of what distortion occurs in practical modulators, in which the rectifiers are not perfect switches, is not so simple to deal with, as all sorts of non-linear laws are encountered. It is clear that the actual law plays a most important part, since Section 4 shows that when square-law

(Footnote continued on next page)

calculation is the analysis recently published by the present author[8] of complex signals applied to a single linear rectifier. Throughout it is assumed that the signal amplitudes are small compared with the carrier.

2. Single Linear Rectifier

It will become clear later that the performance of balanced modulators, however complex, can be described in terms of the equations applying to a single linear rectifier. We will thus examine the intermodulation-distortion of a linear rectifier to which the voltage

$$e = E_1 [\cos pt + x_1 \cos (p - q_1) t + x_2 \cos (p - q_2) t] \quad .. \quad .. \quad (1)$$

is applied; $E_1 \cos pt$ is the carrier, and $x \ll 1$. The analysis will thus cover all ordinary requirements for intermodulation, since as x approaches unity, the distortion will be too large for most applications. Distortion levels around 60 or 70 db below the wanted outputs are commonly specified.

The rectifier has a switching function, $\phi(t)$, generated by the applied voltage, and the output is therefore

$$e \,.\, \phi(t) \quad .. \quad .. \quad .. \quad .. \quad (2)$$

Now $\phi(t)$ is discussed and derived in Reference 8, Equ. (11), and, ignoring the constant term which has no bearing on distortion, is shown to be approximately

$$\phi(t) = h_1 \sum_{n=1}^{\infty} \frac{(-1)^{n-1}}{2n-1} \cos [(2n-1)(pt - x_1 \sin q_1 t - x_2 \sin q_2 t)] \quad (3)$$

This can be expanded into series form with Bessel functions as coefficients; and if for compactness we put $r = 2n - 1$, so that r takes the values 1, 3, 5, etc., we obtain

$$\frac{\phi(t)}{h_1} = \sum \frac{(-1)^{(r-1)/2}}{r} \text{ multiplied by}$$

$$\cos rpt \left[\{J_0(rx_1) + 2J_2(rx_1) \cos 2q_1 t + ..\} \{J_0(rx_2) + 2J_2(rx_2) \cos 2q_2 t + ..\} - \{2J_1(rx_1) \sin q_1 t + 2J_3(rx_1) \sin 3q_1 t + ..\} \{2J_1(rx_2) \sin q_2 t + 2J_3(rx_2) \sin 3q_2 t + ..\}\right] + \sin rpt \left[\{2J_1(rx_1) \sin q_1 t + 2J_3(rx_1) \sin 3q_1 t + ..\} \{J_0(rx_2) + 2J_2(rx_2) \cos 2q_2 t + ..\} + \{2J_1(rx_2) \sin q_2 t + 2J_3(rx_2) \sin 3q_2 t + ..\} \{J_0(rx_1) + 2J_2(rx_1) \cos 2q_1 t + ..\}\right] \quad .. \quad .. \quad .. \quad (4)$$

Now, since $x \ll 1$, and since we are not likely to be concerned with large values of r, we can use the following approximations for the Bessel functions:

$$\left.\begin{aligned} J_0(x) &\approx 1 \\ J_1(x) &\approx x/2 \\ J_2(x) &\approx x^2/8 \\ J_3(x) &\approx x^3/48 \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (5)$$

From these formulae, any desired or undesired output component may be worked out, easily if rather laboriously.

Before listing products of interest, it is as well to point out that there are traps in interpreting the order of intermodulation and order of modulation. A frequency term such as

$$kF \pm mf_1 \pm nf_2$$

is regarded as of the kth order of modulation, and of $(m + n)$th order of intermodulation, but this does not mean that only the terms in $\cos kpt$ and $\sin kpt$ in Equ. (4) contribute to it. Care must be taken to see that all terms which can contribute are included in the calculation. This point is emphasized in the tables of products below. Note that throughout, all output terms are cosines.

The first-order terms are shown in Table 1.

TABLE 1

Product		Relevant values of r in Equ. (4)	Amplitude (Multiply by $h_1 E_1$)
F, f notation	p, q notation		
$F - f_1$	q_1	1	$x_1/2$
$F - f_2$	q_2	1	$x_2/2$
$F + f_1$	$2p - q_1$	1, 3	$x_1/2$
$F + f_2$	$2p - q_2$	1, 3	$x_2/2$

The third-order intermodulation and harmonic terms based on first-order modulation are shown in the first part of Table 2, and those based on third-order modulation are shown in the second part of Table 2. The latter must not be overlooked in practice, as many of them are of considerable importance and relatively large amplitude.

Third-order distortion terms for higher orders of modulation can be readily calculated from the general second- and third-order expansion of Equ. (2) given in Appendix 1. It will be found that the terms $kF \pm 3f$ have an amplitude

$$x^3 \,.\, (-1)^{(k+1)/2} \,.\, h_1 E_1 k/48$$

and the terms $kF \pm f_1 \pm 2f_2$ have an amplitude

rectifiers are used, there is no distortion at all. On the other hand, experience suggests that the distortion magnitudes as calculated for the perfect-switch rectifiers in this paper are not often exceeded in practice, and if an extra margin of say 3–6 db is allowed for effect (a)—it will be seen from Section 9 that this need be allowed only for distortion products of first-order modulation—then designs based on these calculations will not often fail to meet their required performance. At any rate, this will be a much better basis of design than merely to make distortion measurements on one sample modulator and then assume the results obtained will apply to all production modulators of the same nominal design—they will not! Variations from one sample to another may be very large.

$$x_1 x_2^{r} \cdot (-1)^{(k+1)/2} \cdot h_1 E_1 \; k/16.$$

The wanted side-tones, $kF \pm f$, have an amplitude approximately

$$x \cdot (-1)^{(k-1)/2} \cdot h_1 E_1 / 2k$$

The distortion on higher orders of modulation is thus worse by a factor approximately k^2 than that on first-order modulation.

In the single linear rectifier all orders of non-linearity are produced.* For instance, there are second-order terms such as

$$h_1 E_1 \left[-(x_1^2/8) \cos 2q_1 t - (x_2^2/8) \cos 2q_2 t + (x_1 x_2/4) \cos (q_1 - q_2) t - (x_1 x_2/4) \cos (q_1 + q_2) t \right] \quad .. \quad .. \quad (6)$$

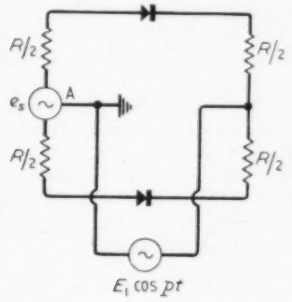

Fig. 1. *Balanced series modulator, where* $e_s = 2E_1 [x_1 \cos (p - q_1) t + x_2 \cos (p - q_2) t]$. A is *the centre point of the balanced signal source.*

TABLE 2

Product		Relevant values of r in Equ. (4)	Amplitude (Multiply by $h_1 E_1$)
F, f notation	p, q notation		
$F - 3f_1$	$2p - 3q_1$	1, 3	$-x_1^3/48$
$F - 3f_2$	$2p - 3q_2$	1, 3	$-x_2^3/48$
$F + 3f_1$	$4p - 3q_1$	3, 5	$-x_1^3/48$
$F + 3f_2$	$4p - 3q_2$	3, 5	$-x_2^3/48$
$F + f_1 + 2f_2$	$4p - q_1 - 2q_2$	3, 5	$-x_1 x_2^2/16$
$F + 2f_1 + f_2$	$4p - 2q_1 - q_2$	3, 5	$-x_1^2 x_2/16$
$F + f_1 - 2f_2$	$q_1 - 2q_2$	1	$-x_1 x_2^2/16$
$F - 2f_1 + f_2$	$2q_1 - q_2$	1	$-x_1^2 x_2/16$
$F - f_1 + 2f_2$	$2p + q_1 - 2q_2$	1, 3	$-x_1 x_2^2/16$
$F + 2f_1 - f_2$	$2p - 2q_1 + q_2$	1, 3	$-x_1^2 x_2/16$
$F - f_1 - 2f_2$	$2p - q_1 - 2q_2$	1, 3	$-x_1 x_2^2/16$
$F - 2f_1 - f_2$	$2p - 2q_1 - q_2$	1, 3	$-x_1^2 x_2/16$
$3F - 3f_1$	$3q_1$	1	$x_1^3/16$
$3F - 3f_2$	$3q_2$	1	$x_2^3/16$
$3F + 3f_1$	$6p - 3q_1$	5, 7	$x_1^3/16$
$3F + 3f_2$	$6p - 3q_2$	5, 7	$x_2^3/16$
$3F + f_1 + 2f_2$	$6p - q_1 - 2q_2$	5, 7	$3x_1 x_2^2/16$
$3F + 2f_1 + f_2$	$6p - 2q_1 - q_2$	5, 7	$3x_1^2 x_2/16$
$3F + f_1 - 2f_2$	$2p - q_1 + 2q_2$	1, 3	$3x_1 x_2^2/16$
$3F - 2f_1 + f_2$	$2p + 2q_1 - q_2$	1, 3	$3x_1^2 x_2/16$
$3F - f_1 + 2f_2$	$4p + q_1 - 2q_2$	3, 5	$3x_1 x_2^2/16$
$3F + 2f_1 - f_2$	$4p - 2q_1 + q_2$	3, 5	$3x_1^2 x_2/16$
$3F - f_1 - 2f_2$	$q_1 + 2q_2$	1	$3x_1 x_2^2/16$
$3F - 2f_1 - f_2$	$2q_1 + q_2$	1	$3x_1^2 x_2/16$

* Note that odd orders of non-linearity are always associated with odd orders of modulation, and even orders of non-linearity with even orders of modulation.

but as all terms produced by even-order non-linearity cancel out in perfectly-balanced modulators, they are not important in the present theoretical study. They can, if desired, be readily written down from the expansion given in Appendix 1.

If intermodulation effects for three or more signal tones are required, they can be deduced in the manner discussed in Reference 8, Section 3.3.

3. Balanced Series Modulator

The working of the previous section can be applied directly to a balanced modulator only if the identical conditions are maintained, namely, that the carrier and signal currents flow through the same circuit, so that their ratio is unaffected by the switching of the rectifiers. Fig. 1 shows a balanced series modulator which meets this requirement; a corresponding shunt modulator is easily devised. In this modulator, one rectifier has an applied e.m.f.

$$e_- = E_1 \left[\cos pt + x_1 \cos (p - q_1) t + x_2 \cos (p - q_2) t \right] \quad .. \quad .. \quad (7)$$

which gives rise to a switching function $\phi_+(t)$, defined by Equ. (3); and the other rectifier has an applied e.m.f.

$$e_- = E_1 \left[\cos pt - x_1 \cos (p - q_1) t - x_2 \cos (p - q_2) t \right] \quad .. \quad .. \quad (8)$$

giving rise to a switching function $\phi_-(t)$ defined by Equ. (3) but with $-x$ replacing x.

The output voltage is evidently

$$e_+ \cdot \phi_+ (t) - e_- \cdot \phi_- (t) \quad .. \quad .. \quad (9)$$

which contains only terms in odd orders of x; i.e., the carrier and all even-order terms are balanced out. The ratio of each of the third-order intermodulation and harmonic terms to the first-order output terms is the same as for the single rectifier.

Note that the signal e.m.f. to the whole modulator is $2xE_1 \cos (p - q) t$ for each frequency,

so that the ratio of (signal voltage)/(carrier voltage) is $2x$ and not x.

4. Use of Square-Law Rectifiers

If the rectifiers in the previous discussion had been square-law instead of linear they would be represented by an output (v_0) to input (v_1) relationship

$$v_0 = a_1 v_1 + a_2 v_1^2 \quad .. \quad .. \quad (10)$$

No distortion products of order higher than the second could then be produced, so that a balanced modulator as shown in Fig. 1, in which all even-order terms are balanced out, would be completely distortionless.

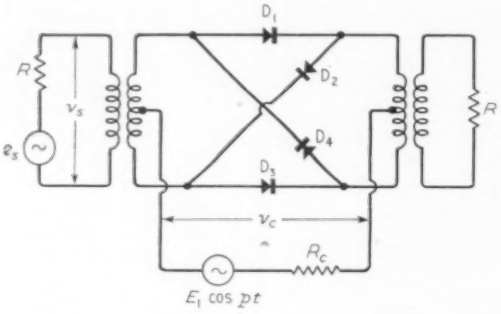

Fig. 2. *Ring modulator, where*

$e_s = 2V_c [x_1 \cos (p = q_1)\, t + x_2 \cos (p - q_2)\, t]$;

$v_c = V_c \cos pt$ — *carrier p.d. when* $e_s = 0$;

v_s = *signal p.d. when modulator is not overloading; e.g. around* $pt = 0, \pi$, *etc.*

5. Ring Modulator with Ideal Rectifiers

The ring modulator with rectifiers having $r_f = 0$ and $r_b = \infty$, and a carrier-generator resistance of zero, was dealt with by Belevitch (for a single signal tone) in his first paper[1] and is a particularly simple case. Fig. 2 shows a ring modulator, and putting in the conditions above, it can be seen that so long as the carrier keeps rectifiers D_1 and D_3 conducting and D_2 and D_4 blocking (or vice versa), the output is the signal voltage v_s. As the carrier approaches zero voltage at $pt = \pi/2$, $3\pi/2$, etc., the condition is reached when the signal and carrier currents through one of the blocking rectifiers become equal and opposite, and then that rectifier starts to conduct and unbalances the circuit. For example, if D_1 and D_3 are conducting and the signal current (assumed altering very slowly) is flowing clockwise, then it is D_2 which starts to conduct as the carrier approaches $pt = \pi/2$. As soon as this happens, the signal is short-circuited and the carrier-balance fails, so that the output is then the carrier voltage v_c (i.e., $v_c/2$ is produced across half the primary of the output transformer, giving v_c across the secondary). This condition continues until at $v_c = 0$, D_3 ceases to conduct and D_4 ceases to block, but this change does not alter the output, which is still v_c. Then, as v_c increases negatively, a point is reached when the carrier current in D_1 becomes equal and opposite to the signal current, and then D_1 ceases to block; the output then becomes v_s again, but reversed in phase, and there is no output due to the carrier, since the circuit is once more balanced.

The output voltage can thus be written as

$$v_s [\phi_+(t) + \phi_-(t)] + v_c [\phi_+(t) - \phi_-(t)] \quad .. \quad (11)$$

where $\phi(t)$ is as defined by Equ. (3) with subscripts as used in Equ. (9) and with $h_1 = 2/\pi$.

The way in which this expression represents the circuit action as described above will probably be easier to see with the help of Fig. 3, which shows the waveforms of the switching functions, assuming the signal voltage changes only slowly; i.e., is of relatively low frequency. In these functions, x is the ratio of signal p.d. to carrier p.d. (i.e., v_s/v_c) since this is the ratio of signal current to carrier current through the rectifier which fails, both before and after it switches over.

Equ. (11) can be rewritten as

$$e_+ \,.\, \phi_+(t) - e_- \,.\, \phi_-(t) \quad .. \quad .. \quad (12)$$

using the symbolism of Equs. (7) and (8), and this output is identical with that of the balanced series modulator, Equ. (9), except for the value of h_1 which is used in determining $\phi(t)$. This means that although the ring modulator is more efficient, its distortion ratios are identical with those of the previous circuit, provided the (signal voltage)/(carrier voltage) ratio is x instead of $2x$ as used in the previous circuit.

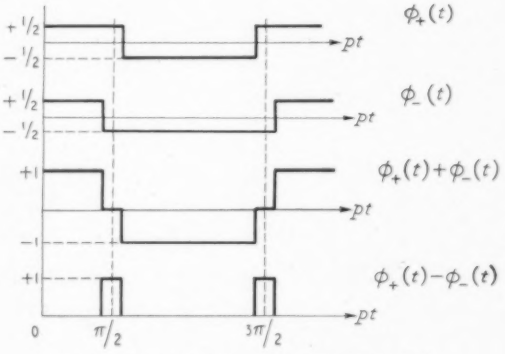

Fig. 3. *Switching functions in ring modulator.*

It should be noted that Belevitch calculated the $F \pm 3f$ and $3F \pm 3f$ products, and the values given here agree exactly with his, provided x is small.

6. Ring Modulator with Non-Ideal Rectifiers

If the rectifiers have finite forward and backward resistances and if the carrier generator resistance is not zero, the analysis of the circuit is greatly complicated. It follows the lines of the previous section, but as (taking the previous example) when D_1 and D_2 are both conducting they do not now completely short-circuit the signal voltage, there is a residual signal voltage opposing the carrier across D_3. Thus D_3 ceases to conduct before v_c reaches zero. Similarly D_4 does not start to conduct until after v_c has become negative. There are thus four points of discontinuity per half-cycle of carrier instead of two, and thus the equation for the output voltage involves two functions like Equ. (12) instead of one, and with coefficients involving r_f, r_b, R and R_c.

Belevitch[3] analyses this case in detail for one signal tone, and his analysis can be extended to two or more signal components by replacing his switching ('signature') functions by the corresponding $\phi(t)$ functions.* But he shows that the difference between the results obtained and those of the ideal modulator (Section 5) is very small in practical cases—generally less than 1 db—and therefore the full analysis is hardly worth while.

7. Shunt and Series Modulators of Conventional Type

The modulator of Fig. 1 was, of course, a special arrangement designed to fulfil a particular requirement of the analysis. Conventional shunt and series modulators do not use a balanced signal source with the carrier returned to its centre point, and they generally have a finite carrier-generator resistance. When they are used in pairs with a common carrier supply, connected in parallel opposition, or when their carrier generator resistance is very small, they do give much the same intermodulation performance as ring modulators. But when used singly, the effect of the carrier-generator resistance is very large; and it is sometimes possible to find a value which gives a minimum distortion 20 or 30 db below that obtained with zero resistance. This was originally described and explained by Tucker and Jeynes[2] and analysed by Belevitch[3,4] Belevitch's analysis can be extended to two or more signal components as previously discussed for the ring modulator.

8. Use of a Square-Wave Carrier Voltage

Since the non-linear effects described above are due to the interference of the signal voltage with the switching action, which is nominally under the control of a sinusoidal carrier e.m.f., it is clear that these effects can be eliminated by using a square-wave carrier provided the signal voltage never exceeds the carrier.

It is not, of course, always convenient or economical to use a square-wave carrier; this applies particularly in carrier telephony, although there are doubtless many special applications where the extra complication would be well justified. But even when the square-wave is used, there is the residual non-linear distortion due to the non-linearity of the rectifier forward and backward resistances. The amount of this distortion is quite unrelated to that due to switching, but measured results given by Christiansen[6] for ring modulators using germanium rectifiers with a square-wave carrier show distortion of the same order as, or slightly higher than, has been calculated for the switching effect. It is thus evident that the use of a square-wave carrier will not necessarily reduce distortion appreciably, and it is necessary in any particular case to investigate the relative magnitudes of the two effects before the trouble and expense of providing a square-wave carrier is decided upon. Furthermore, it is probable that the calculated distortion for the switching effect is somewhat approximate since practical rectifiers do not switch suddenly at zero voltage as has been assumed.

Christiansen shows that the non-linearity of the rectifier forward resistance can be reduced by the use of a resistor in series with each rectifier. Usually an optimum value can be found giving the minimum level of distortion products relative to the wanted components. Negative feedback can also improve this form of distortion.

9. Equivalent Circuits of Ring Modulator with Non-Linear Response

The ring modulator in its linear mode (i.e., all non-linearity of any kind either ignored or made negligible by the use of low signal/carrier voltage ratio) can be represented by a time-invariable lattice followed by a perfect commutator operating at the carrier frequency. The lattice has one pair of arms with a resistance r_f and the other with a resistance r_b; and it is assumed that the rectifiers from which they are derived switch over at zero applied voltage.

Non-linearity of the rectifier forward and backward resistances can be included in this circuit by making the lattice arms voltage-dependent, as shown in Fig. 4, and we may allow for third-order distortion by saying that $r_f(v)$ and $r_b(v)$ are such that the lattice response can be represented by the cubic

$$v_R = a_1 v_S + a_3 v_S^3 \quad .. \quad .. \quad (13)$$

* Note that $2\phi(t)$ corresponds to the signature function.

On substituting the input signal expression for v_s, the Fourier expansion of v_R is obtained; and then all the components of v_R are multiplied by the switching function to obtain the series for the output components. This means that all orders of modulation have the same relationships between their first-order sidetones and their third-order intermodulation and harmonic products; actual levels are proportional to $1/k$ (where k is the order of modulation).

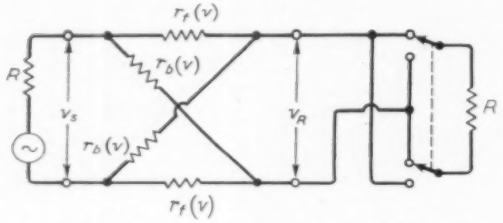

Fig. 4. *Equivalent circuit of ring modulator.*

In the case of third-order non-linear distortion due to the switching-interference effect analysed in this paper, this simple equivalent circuit does not apply, since the relationships within the various orders of modulation are different, and the effect is dependent on the ratio of the signal voltage to the carrier, instead of on its actual magnitude. But since in each order of modulation, the relation between the third-order intermodulation and harmonic terms, and the effect of the third-order distortion on the amplitude of the first-order sidetones, are the same as in Equ. (13), the equivalent circuit of Fig. 4 can be used by means of an artifice. This consists of taking different coefficients in Equ. (13) for each order of modulation and of using x_s in place of v_s. Thus

$$v_R = E_1 (b_1 x_s + b_3 x_s^3) \quad \ldots \quad \ldots \quad (14)$$

where

$$b_3 = -k^2 b_1/6 \text{ and } x_s = v_s/E_1 \quad \ldots \quad (15)$$

the appropriate value of k being used in calculating the distortion due to the kth order of modulation.

The principle of the equivalent circuit appears to apply to all orders of distortion. Its general application is discussed in Appendix 2.

10. Conclusions

The main conclusions from the work discussed above have been given in the Summary at the head of the paper.

Acknowledgments

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APPENDIX 1

The Complete Output Expression for First-Order Sidetones and Second- and Third-Order Distortion Products from a Single Linear Rectifier with a Large Carrier and Two Small Signal Tones Applied

In the expression below, all terms involving x in higher degree than the third have been neglected. The amplitude of any required frequency component can readily be determined by substituting the appropriate one or two values of r in the relevant output terms.

Output of single linear rectifier with complex signal $E_1\{\cos pt + x_1 \cos(p-q_1)t + x_2 \cos(p-q_2)t\}$ applied

$$= h_1 E_1 \sum_{r=1,\,3,\,5 \text{ etc.}}^{\infty} (-1)^{(r-1)/2} \text{ of:—}$$

$$\left\{\frac{1}{2r} - x_1^2\left(\frac{1}{4}+\frac{r}{8}\right) - x_2^2\left(\frac{1}{4}+\frac{r}{8}\right)\right\}\cos(r+1)pt$$

$$+\left\{\frac{1}{2r} + x_1^2\left(\frac{1}{4}-\frac{r}{8}\right) + x_2^2\left(\frac{1}{4}-\frac{r}{8}\right)\right\}\cos(r-1)pt$$

$$+\left\{-\frac{x_1}{4} + x_1x_2^2\left(\frac{r}{8}+\frac{r^2}{16}\right) + x_1^3\left(\frac{r}{16}+\frac{r^2}{32}\right)\right\} \cos\{(r+1)p+q_1\}t$$

$$+\left\{-\frac{x_1}{4} + \frac{x_1}{2r} - x_1x_2^2\left(\frac{r}{4}-\frac{r^2}{16}\right) - x_1^3\left(\frac{r}{8}-\frac{r^2}{32}\right)\right\} \cos\{(r-1)p+q_1\}t$$

$$+\left\{x_1\left(\frac{1}{4}+\frac{1}{2r}\right) - x_1x_2^2\left(\frac{r}{4}+\frac{r^2}{16}\right) - x_1^3\left(\frac{r}{8}+\frac{r^2}{32}\right)\right\} \cos\{(r+1)p-q_1\}t$$

$$+\left\{\frac{x_1}{4} + x_1x_2^2\left(\frac{r}{8}-\frac{r^2}{16}\right) + x_1^3\left(\frac{r}{16}-\frac{r^2}{32}\right)\right\} \cos\{(r-1)p-q_1\}t$$

+ terms of the form $\cos\{(r \pm 1)p \pm q_2\}t$ having amplitudes as above but with x_1 replaced by x_2 and vice-versa.

$$+\frac{rx_1^2}{16}\cos\{(r+1)p+2q_1\}t$$

$$+x_1^2\left(\frac{r}{16}-\frac{1}{4}\right)\cos\{(r-1)p+2q_1\}t$$

$$+x_1^2\left(\frac{r}{16}+\frac{1}{4}\right)\cos\{(r+1)p-2q_1\}t$$

$$+\frac{rx_1^2}{16}\cos\{(r-1)p-2q_1\}t$$

+ similar terms of the form $\cos\{(r \pm 1)p \pm 2q_2\}t$ as above

$$\left.\begin{aligned} &-\frac{r^2x_1^3}{96}\cos\{(r+1)p+3q_1\}t \\ &+\left(\frac{rx_1^3}{16}-\frac{r^2x_1^3}{96}\right)\cos\{(r-1)p+3q_1\}t \\ &+\left(\frac{rx_1^3}{16}+\frac{r^2x_1^3}{96}\right)\cos\{(r+1)p-3q_1\}t \\ &+\frac{r^2x_1^3}{96}\cos\{(r-1)p-3q_1\}t \end{aligned}\right\}$$

+ similar terms of the form $\cos\{(r \pm 1)p \pm 3q_2\}t$

$$+\frac{r x_1 x_2}{8}\cos\{(r+1)p+q_1+q_2\}t$$
$$+x_1x_2\left(\frac{r}{8}-\frac{1}{2}\right)\cos\{(r-1)p+q_1+q_2\}t$$
$$+x_1x_2\left(\frac{r}{8}+\frac{1}{2}\right)\cos\{(r+1)p-q_1-q_2\}t$$
$$+\frac{r x_1 x_2}{8}\cos\{(r+1)p-q_1-q_2\}t$$
$$-x_1x_2\left(\frac{r}{8}+\frac{1}{4}\right)\cos\{(r+1)p+q_1-q_2\}t$$
$$+x_1x_2\left(-\frac{r}{8}+\frac{1}{4}\right)\cos\{(r-1)p+q_1-q_2\}t$$
$$-x_1x_2\left(\frac{r}{8}+\frac{1}{4}\right)\cos\{(r+1)p-q_1+q_2\}t$$
$$+x_1x_2\left(-\frac{r}{8}+\frac{1}{4}\right)\cos\{(r-1)p-q_1+q_2\}t$$

$$-\frac{r^2x_1^2x_2}{32}\cos\{(r+1)p+2q_1+q_2\}t$$
$$+x_1^2x_2\left(\frac{3r}{16}-\frac{r^2}{32}\right)\cos\{(r-1)p+2q_1+q_2\}t$$
$$+x_1^2x_2\left(\frac{r}{16}+\frac{r^2}{32}\right)\cos\{(r+1)p+2q_1-q_2\}t$$
$$+x_1^2x_2\left(-\frac{r}{8}+\frac{r^2}{32}\right)\cos\{(r-1)p+2q_1-q_2\}t$$
$$-x_1^2x_2\left(\frac{r}{8}+\frac{r^2}{32}\right)\cos\{(r+1)p-2q_1+q_2\}t$$
$$+x_1^2x_2\left(\frac{r}{16}-\frac{r^2}{32}\right)\cos\{(r-1)p-2q_1+q_2\}t$$
$$+x_1^2x_2\left(\frac{3r}{16}+\frac{r^2}{32}\right)\cos\{(r+1)p-2q_1-q_2\}t$$
$$+\frac{r^2x_1^2x_2}{32}\cos\{(r-1)p-2q_1-q_2\}t$$

$$-\frac{r^2x_1x_2^2}{32}\cos\{(r+1)p+q_1+2q_2\}t$$
$$+x_1x_2^2\left(\frac{3r}{16}-\frac{r^2}{32}\right)\cos\{(r-1)p+q_1+2q_2\}t$$
$$+x_1x_2^2\left(\frac{r}{16}+\frac{r^2}{32}\right)\cos\{(r+1)p-q_1+2q_2\}t$$
$$+x_1x_2^2\left(-\frac{r}{8}+\frac{r^2}{32}\right)\cos\{(r-1)p-q_1+2q_2\}t$$
$$-x_1x_2^2\left(\frac{r}{8}+\frac{r^2}{32}\right)\cos\{(r+1)p+q_1-2q_2\}t$$
$$+x_1x_2^2\left(\frac{r}{16}-\frac{r^2}{32}\right)\cos\{(r-1)p+q_1-2q_2\}t$$
$$+x_1x_2^2\left(\frac{3r}{16}+\frac{r^2}{32}\right)\cos\{(r+1)p-q_1-2q_2\}t$$
$$+\frac{r^2x_1x_2^2}{32}\cos\{(r-1)p-q_1-2q_2\}t$$

APPENDIX 2

The General Application of the Equivalent Circuit of Fig. (4) *and Section* 9

It is shown below that for any order of modulation (k) and for any order of harmonic distortion (m), the equivalent circuit, in which a non-linear lattice precedes a perfect switch (i.e., commutator), holds exactly for switching-interference provided a different set of coefficients is taken for each order of modulation. These coefficients are those of the power series

$$v_R = b_1v_s + b_3v_s^3 + \ldots. + b_n v_s^n + \ldots. \qquad (1)$$

and the different set for each order of modulation is conveniently indicated by a prefix; e.g., ${}_kb_n$ is the coefficient of v_s^n when the kth order of modulation is being considered. The proof is based on the expressions for harmonic distortion in an ideal ring modulator obtained by Belevitch[1], with the correction of a misprint and the addition of correct signs obtained from the paper by Bennett[3] which he quotes. It does not seem possible to extend this proof to intermodulation products without a quite prohibitive amount of labour; but as the circuit is shown in Section 9 to be valid for third-order intermodulation, and is valid for all orders of harmonic distortion, it is not unreasonable to conclude (without further proof) that it is also valid for all orders of intermodulation distortion.

The independent variable is shown in (1) as a voltage, but this is only appropriate if the carrier voltage $E_1 = 1$. In general we should write

$$v_R = E_1[b_1x + b_3x^3 + \ldots. + b_nx^n + \ldots.] \qquad (2)$$

since the non-linearity due to switching interference is dependent on the ratio of signal/carrier voltage, as distinct from the rectifier non-linearity which is, at least mainly, dependent on the signal voltage itself. This is a further point of artificiality about the equivalent circuit, although it gives a means of separating the two effects in practical measurements.

The applied signal is a single tone $xE_1\cos\theta$, where $\theta = (p-q)t$ or $2\pi ft$. Let the Fourier series for v_R be

$$E_1[c_1\cos\theta + c_3\cos 3\theta + c_5\cos 5\theta + \ldots. + c_r\cos r\theta + \ldots.] \qquad (3)$$

To relate this to the power series (2), we use the expansion

$$\cos^n\theta = \frac{1}{2^n}\left[\cos n\theta + n\cos(n-2)\theta + \frac{n(n-1)}{2!}\cos(n-4)\theta + \ldots. + \frac{n!}{n!}\cos(n-2n)\theta\right] \qquad (4)$$

which is more conveniently written for odd values of n as

$$\cos^n\theta = \frac{1}{2^{n-1}}\left[\cos n\theta + n\cos(n-2)\theta + \frac{n(n-1)}{2!}\cos(n-4)\theta + \ldots.\right] \qquad (5)$$

where the series terminates at $\cos\theta$.

From this it is easy to derive the value of c_r in terms of b, thus:

$$c_r = \frac{1}{2^{r-1}}\left[b_rx^r + \frac{1}{2^2}(r+2)\,b_{r+2}x^{r+2} + \frac{1}{2^4}\cdot\frac{(r+4)(r+3)}{2!}b_{r+4}x^{r+4} + \ldots.\right] \qquad (6)$$

or

$$c_r = \frac{1}{2^{r-1}}\sum_{\gamma=0}^{\infty}\frac{1}{2^{2\gamma}}\cdot\frac{(r+2\gamma)!}{\gamma!\,(r+\gamma)!}\cdot b_{r+2\gamma}\cdot x^{r+2\gamma} \qquad (7)$$

and the values of the b coefficients can be determined by taking special values of the c coefficients, each being that obtained by making the series (3) terminate at it. Thus if the series is made to terminate at the rth term, then from (5) or (6),

$$b_r = 2^{r-1}c_r'/x^r \qquad (8)$$

where c_r' is the special value of c_r.

We are now in a position to test the validity of this for distortion due to switching interference. Belevitch (corrected) shows that the output voltage of an ideal ring modulator, of the kth order of modulation and the mth order of harmonic distortion, is

$$\frac{V_{k,m}}{E_1} = (-1)^{[(k+m)/2]+1} \cdot x^m \frac{\Gamma\left(\frac{k+m-1}{2}\right)}{\pi m!\,\Gamma\left(\frac{k-m+3}{2}\right)} \cdot {}_2F_1 \quad .. \quad .. \quad .. \quad (9)$$

where ${}_2F_1$ is the Gaussian hypergeometric function

$$F\left(\frac{k+m-1}{2},\ \frac{m-k-1}{2};\ m+1;\ x^2\right)$$

which can be expanded, if $x<1$, as

$${}_2F_1 = 1 + \frac{(k+m-1)(m-k-1)}{4(m+1)}x^2 + \frac{(k+m-1)(k+m+1)(m-k-1)(m-k+1)}{16(m+1)(m+2)\,.\,2!}x^4 + \ldots . \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (10)$$

Now if we obtain values of b_r as in (8) by using (9) with ${}_2F_1$ put equal to unity, a test of the validity of the equivalent circuit is to see if it then gives the same value of $V_{k,m}$ as obtained from (9) fully expanded. Thus, remembering that the output of the equivalent circuit is v_R followed by a perfect commutator [i.e., the output of kth order of modulation and mth order of harmonic distortion is $2E_1\, c_m\, (-1)^{(k-1)/2}/\pi k$] we say that

$${}_kc_m' = \frac{k}{2}(-1)^{(m-1)/2} \cdot \frac{\Gamma\left(\frac{k+m-1}{2}\right)}{m!\,\Gamma\left(\frac{k-m+3}{2}\right)} \cdot x^m \quad (11)$$

So that*

$${}_kb_m = 2^{m-2}\,(-1)^{(m-1)/2} \frac{\Gamma\left(\frac{k+m-1}{2}\right)}{m!\,\Gamma\left(\frac{k-m+3}{2}\right)} \cdot k \quad (12)$$

Using this, the complete value of ${}_kc_m$ is, from (7),

$${}_kc_m = \frac{k}{2^{m-1}} \sum_{\gamma=0}^{\infty} \frac{1}{2^{2\gamma}} \cdot \frac{(m+2\gamma)!}{\gamma!(m+\gamma)!} \cdot 2^{m-2+2\gamma}\,(-1)^{[(m-1)/2]+\gamma} \cdot \frac{\Gamma\left(\frac{k+m-1+2\gamma}{2}\right)}{(m+2\gamma)!\,\Gamma\left(\frac{k-m+3-2\gamma}{2}\right)} x^{m+2\gamma} \quad .. \quad (13)$$

$$= (-1)^{(m-1)/2} \cdot \frac{k}{2} \cdot x^m \sum_{\gamma=0}^{\infty} (-1)^{\gamma} \cdot \frac{\Gamma\left(\frac{k+m-1+2\gamma}{2}\right)}{\Gamma\left(\frac{k-m+3-2\gamma}{2}\right)} \cdot \frac{x^{2\gamma}}{\gamma!\,(m+\gamma)!} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (14)$$

and noting that

$$\frac{\sum_{\gamma=0}^{\infty} \Gamma\left(\frac{m-k-1+2\gamma}{2}\right)}{\Gamma\left(\frac{k-m+3}{2}\right)\Gamma\left(\frac{m-k-1}{2}\right)} = \sum_{\gamma=0}^{\infty} \frac{(-1)^{\gamma}}{\Gamma\left(\frac{k-m+3-2\gamma}{2}\right)} \quad .. \quad .. \quad .. \quad (15)$$

and introducing a factor $m!\,\Gamma\left(\frac{k+m-1}{2}\right)$ in both numerator and denominator, (14) becomes

$${}_kc_m = \frac{k}{2}(-1)^{(m-1)/2} \cdot x^m \cdot \frac{\Gamma\left(\frac{k+m-1}{2}\right)}{m!\,\Gamma\left(\frac{k-m+3}{2}\right)} \sum_{\gamma=0}^{\infty} \frac{\Gamma\left(\frac{k+m-1+2\gamma}{2}\right)\Gamma\left(\frac{m-k-1+2\gamma}{2}\right) . m!}{\Gamma\left(\frac{k+m-1}{2}\right)\Gamma\left(\frac{m-k-1}{2}\right) .\ (m+\gamma)!\ \gamma!} x^{2\gamma} \quad .. \quad (16)$$

and on multiplying this by $2E_1\,(-1)^{(k-1)/2}/\pi k$ to obtain the actual voltage from the complete equivalent circuit including the commutator, we find that this is exactly the same as $V_{k,m}$ given by (9). [Note that the expression following the summation sign in (16) is the generalized form of the hypergeometric function ${}_2F_1$.]

This proves that the equivalent circuit of Fig. 4, used as described, is entirely self-consistent in respect of harmonic-distortion.

For numerical evaluation, it should be remembered that $\Gamma(\frac{1}{2}) = \sqrt{\pi}$, $\Gamma(-\frac{1}{2}) = -2\sqrt{\pi}$, and that for any variable z,

$$\Gamma(z+1) = z\Gamma(z) = z(z-1)\Gamma(z-1) \text{ etc.}$$

Any intermodulation product can now be rapidly determined, since it is easy to prove from the power series by induction that products, $kF \pm mf_1 \pm nf_2$, of the $(m+n)$th order of intermodulation, have an amplitude given to the first order of accuracy (still assuming $x \ll 1$) by

$$\frac{2(-1)^{(k-1)/2}E_1}{\pi k} \cdot {}_kb_{m+n} \cdot \frac{(m+n)!}{2^{m+n-1}\,m!\,n!}\, x_1^{\,m} x_2^{\,n} \quad .. \quad (17)$$

where ${}_kb_{m+n}$ is determined from Equ. (12).

* Note that ${}_kb_1$ is unity, whatever the value of k, and that ${}_kb_3 = -k^2/6$. In general, ${}_kb_m$ is a polynomial in k of the $(m-1)$th degree.

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AUTOMATIC TUNING FOR PRIMARY RADAR

By S. Ratcliffe, B.Sc.

(*Continued from p. 131, May issue*)

Complete A.F.C. Loop

For the purposes of the following discussion, radar a.f.c. systems will be divided into four categories:—

(a) Systems in which the change in oscillator frequency is controlled by some non-linear circuit whose behaviour is determined by the sign of the tuning error indicated by the discriminator, and not by its magnitude (once the error exceeds some low value).

(b) Systems in which the change in oscillator frequency is proportional to the magnitude of the tuning error, as well as its sign, and in which the change in oscillator frequency occurs during the interval between transmitter pulses. It will be assumed that the change in oscillator frequency during the transmitter pulse itself is negligible.

(c) Systems similar to those in (b) but in which the frequency correction is actually applied during the pulse, and in which some 'clamp' circuit is then used to hold the tuning fixed at this point until the transmitter next fires.

(d) Hybrid systems which use some combination of the three methods listed above.

For the moment, it will be assumed that such tuning errors as may occur do not exceed the limits of the linear portion of the discriminator curve.

The a.f.c. systems mentioned under (a) above, contain some form of circuit having a small number of stable states, usually two. It is arranged that the oscillator frequency drifts in a direction determined by the state of this circuit and at a rate determined by the circuit constants.

A simple example of this type of circuit is the 'gas-tube' a.f.c.,[7] illustrated in Fig. 13. The oscillator frequency is controlled by variation of the reflector voltage. When in the neighbourhood of the desired frequency, the reflector, which is connected to a capacitor C_1 charging through a resistance R_1, is allowed to drift steadily more positive until the desired frequency is passed. A positive output pulse is then received from the discriminator which causes a gas triode to fire. This gas triode then discharges a capacitor C_2 which is connected through R_2 to the larger capacitor C_1, feeding the reflector. Having discharged C_2, the gas triode is extinguished; C_2 then starts to recharge to the reflector potential, lowering the potential across C_1 in the process. This lowering of the reflector potential normally changes the sign of the tuning error, so that the gas triode will remain extinguished until the charging of the reflector network carries the oscillator frequency back through the desired value. Thus, in operation, the gas triode fires at such a rate as to keep the mean frequency of the klystron at the desired value; any change in the transmitter frequency being accompanied by a change in the rate at which the gas triode fires.

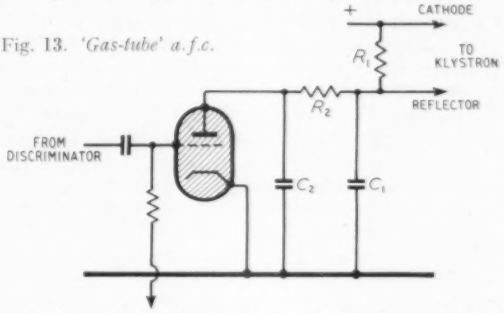


Fig. 13. *'Gas-tube' a.f.c.*

A detailed analysis of this circuit is published elsewhere,[7] and for purposes of the present discussion a simplification of the problem will suffice. Suppose that the sensitivity of the discriminator is infinite; that is, that an infinitesimal error will suffice to produce a voltage large compared with the grid base of the gas triode. Suppose also that the reflector voltage drifts up or down at the same rate (p Mc/s per pulse, say) according to the sign of the tuning error. Then the excursions of the oscillator from the desired value will have a peak amplitude of $\pm p$ Mc/s where p is the rate of follow of the a.f.c. system in Mc/s per pulse.

In practice, there will be a threshold level of error below which the signal from the discriminator will be too small to trigger the gas triode, it is impossible for the circuit designer to maintain p at the desired value over the working range of the reflector voltages, and further variations arise from tolerances on the gas-triode characteristics. A typical design may well have a minimum rate of follow (Mc/s per pulse) which is no higher than 0·05 of the maximum tuning error, so that an a.f.c. of this type must be very slow or inaccurate or both.

Another application of this type of control is the 'bang-bang' control system used with thermally-tuned oscillators. Its performance is not appreciably better than the figure quoted above for the 'gas-tube' system.

No general analysis of systems of this type will be attempted here although McColl[11] has indicated one possible approach. In general, an *ad hoc* argument will suffice to reveal the salient characteristics of any given system and, in fact, the ease with which the performance can be analysed is one of the major attractions of this type of circuit.

One further example[12] of a control system of this type may be cited. This uses mechanical tuning of the oscillator. The tuning shaft is coupled through a differential gear to two 'stepping' motors, similar to those used to drive a uniselector, the two mechanisms being arranged to drive the tuner shaft in opposite directions. The two motor windings are controlled by relays in the anode circuits of a pair of valves fed in push-pull from the discriminator output. If the tuning error, in either direction, exceeds some threshold value, one of the relays will close, and the tuning shaft will move by one step in the required direction. The process will then repeat until the error has fallen below the threshold value. This control system can thus be regarded as having three stable states, corresponding to positive, negative, and negligible error. The stability condition for this system is that a single step of the motor must never cause the tuning to change by twice the threshold level, otherwise the desired central state will not be achieved. The rate of follow of this system is then fixed by the speed of the motor in steps per second.

It will be seen that there are inherent limitations on the speed and locking accuracy of control systems of this type, and a better performance is possible if full use is made of the information available from the discriminator as to the relative magnitude of the tuning error.

The control systems most commonly used in radar a.f.c. fall into category (b) of the classification given above. The pulses from the discriminator are 'stretched' in some way, and the resulting error voltage, possibly after some further 'smoothing', controls the oscillator frequency. In a reflector a.f.c. system, for example, the reflector is often connected to the anode of a Miller integrator, into the grid of which is fed a current proportional to the error voltage. The frequency, therefore, drifts in a suitable direction at a rate proportional to the tuning error existing at the time; or, more correctly, at a rate proportional to the error voltage output from the discriminator or smoothing circuit.

The above account of the operation could, perhaps with some change in nomenclature, equally well have referred to any of the other servo-mechanisms which are sometimes associated with a radar, a lock-following strobe for example. It is not, however, commonly necessary or possible for such systems to move their output shaft, or its equivalent, any significant distance between one pulse and the next, so that servos are commonly designed by negative-feedback theory; i.e., as though error information were continuously available from the discriminator, the 'undesirable' modulation of this voltage at the p.r.f. being removed by suitable filtering.[13]

If we are to attempt an a.f.c. system which is to deal with scanner pulling, we require an approach to the design problem which takes account of the limitation on the rate at which information is available from the discriminator, since the transmitter frequency may change appreciably between one pulse and the next. In a reflector a.f.c. system, for example, this may be the only important limitation on the response speed of the system. A system limited in this way by the supply of error information forms what McColl[11] terms a 'sampling' servo-mechanism, and its analysis is more complicated than that of one with continuous feedback.

If the tuning error in an a.f.c. system is measured relative to the cross-over point on the discriminator characteristic, there are two types of error that can arise. The first of these is a static error that will persist even if the transmitter frequency remains constant for a long period. When the steady-state is reached, the fact that the error voltage must be derived from pulses at finite intervals of time is not important, and the steady-state error e_s will be given by normal negative-feedback theory as

$$e_s = E/(1 + G) \qquad .. \qquad .. \qquad (3)$$

where E is the error that would exist in the absence of a.f.c. feedback, and G can be defined as the ratio of the change in local-oscillator frequency, due to the operation of the a.f.c., to the tuning error which produces the a.f.c. effect. There is usually no difficulty in providing sufficient gain to reduce this static error to negligible proportions, and it will be neglected in discussing the more serious dynamic errors which arise when the transmitter frequency is subject to variation.

The problem of determining the dynamic response and stability conditions for a sampling servo is discussed in some detail in the Appendix to this article. A brief non-mathematical examination of the problem is given here.

The a.f.c. is given a piece of information about the tuning error each time the transmitter fires, and must arrange to take some suitable action after each pulse with a view to reducing the error when the transmitter next fires. In other words,

the problem is to extrapolate from some given data. The a.f.c. will base its extrapolation on the most recent pulse and possibly on some earlier ones, the number of pulses used for this purpose being fixed by the number of independent elements (time-lags) in the feedback loop which are capable of 'remembering' a voltage from one pulse to the next.

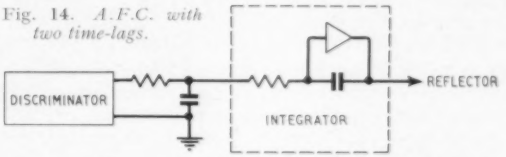
This statement perhaps requires some justification. Consider, for example, the circuit shown in Fig. 14 which possesses two 'time-lags'—both capacitance-resistance networks. The voltage across the first capacitor will depend on all the error voltages which have ever appeared across the discriminator output, so that it may appear that the a.f.c. will base its behaviour at any instant on all the previous error pulses and not merely on the two most recent. However, the response at any instant is completely determined by the voltages on the two capacitors, and these two voltages can be determined from a knowledge of the last two errors. Thus, although previous errors may have been responsible for the building-up of charges on these two capacitors, the behaviour of the a.f.c. is completely determined if the last two errors are known. Hence the statement that the behaviour depends only on these two errors.

If the a.f.c. is to follow a smooth variation in the transmitter frequency which takes place over a period of many pulses, the more of these pulses taken into account in extrapolating, the greater will be the possible accuracy with which the extrapolation can be performed. It has been shown, however, when discussing scanner pulling, that in practice the a.f.c. is required to deal with a more or less random fluctuation in frequency from one pulse to the next, and under these circumstances the a.f.c. will be slowed down by any time-lags which retain information about pulses other than the most recent, just when the most rapid reaction is required.

If the a.f.c. is to extrapolate with reasonable success, it is essential that it should have an appropriate rate of response to an error. It is shown in the Appendix that a single time-lag extrapolating a.f.c., given optimum design, has a response time-constant which may be as great as $R/2$ pulse intervals, where R is the ratio of the maximum to the minimum possible gain round the a.f.c. loop. Thus a good transient response can only be guaranteed if R is kept low. Similar results hold for systems having more than one time-lag. Although it is possible to devise circuits which to some extent compensate for tolerances on gain when following a smoothly-varying transmitter frequency, no circuit has been devised which has the same effect when faced with a rapid and random variation in frequency.

Pending detailed discussion of the problem of reducing the tolerances contributing to R, it can be taken that a value of 6–8 represents the minimum that is obtainable in any type of control system.

Fig. 14. *A.F.C. with two time-lags.*



The third type of a.f.c. system defined above was invented by C. Baron, of R.R.E., in an attempt to avoid the fundamental limitations of the extrapolating systems. In this system, often termed an 'instantaneous' a.f.c., the feedback loop is given sufficiently short time-constants to enable the device to operate during the transmitter pulse as a true negative-feedback system, and a 'clamp' circuit is then used to hold the reflector potential constant during the interval between pulses. An example of a practical circuit of this type is given later, and the Appendix discusses the theory of this system. Except possibly for very short pulses, the design does not present any major difficulty; analysis shows that, although the performance is still determined by the tolerances, it is less sensitive to them than is the extrapolating a.f.c. The classification of linear a.f.c. systems as either 'extrapolating' or 'instantaneous' is convenient, but the boundary line is not necessarily clear. Given a sufficiently wide pulse, an a.f.c. designed as an extrapolating system may, in practice, make an appreciable correction to the frequency during the pulse, so that the system becomes some sort of hybrid.

Throughout the preceding discussion it has been assumed that the tuning error lies within the peaks of the discriminator curve. An a.f.c. is capable of 'holding-in' so long as the tuning error, after the application of the correction, lies within the discriminator peaks. In the absence of the a.f.c. correction, the tuning error may be more than 10 times as great as the discriminator peak separation, and for an uncorrected tuning error of this magnitude the discriminator output will have fallen virtually to zero, so that the a.f.c. will be unable to deal with this error once it has developed. It is, therefore, necessary to distinguish between the a.f.c. 'hold-in' range, which is usually determined by the electronic-tuning range of the klystron, and the 'pull-in' range which gives the limits of initial tuning error with which the a.f.c. can cope. This latter is usually about twice the discriminator peak separation, though it may be greater if the discriminator has been designed to have large skirts to its response curve. Unless

some additional steps are taken, therefore, the a.f.c. cannot be relied upon to deal with frequency drifts in excess of the pull-in range. It is, therefore, customary to provide some means of searching over the electronic-tuning range of the oscillator for the wanted signal, both on switching-on and on any later occasion, this search usually being stopped automatically when signals are received from the discriminator.

Earlier reflector a.f.c. technique provided preset adjustments of the mechanical tuning and of the potential applied to the reflector in the absence of a voltage from the discriminator. The search circuit has the added advantage that it reduces the accuracy with which these presets need be adjusted. It must be pointed out that it is not necessary to use electronic tuning for this search, since by means of the mechanical tuner one can search a much wider band of frequencies, if this should be necessary. For the moment, it will be assumed that both the search and the a.f.c. loop use electronic tuning, discussion of the other possibilities being left till later.

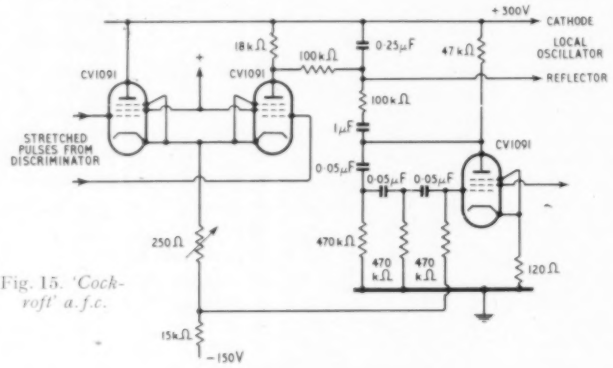


Fig. 15. *'Cockroft' a.f.c.*

The search of the electronic-tuning range is usually performed by applying a 2–6-c/s sawtooth or sine wave to the reflector. Two types of circuit have been used for this purpose.

The first of these is exemplified by the 'Cockroft' circuit[14] shown in outline in Fig. 15. This searches by applying a fraction of the output of a 4-c/s phase-shift oscillator to the reflector. The two discriminator diodes develop negative outputs, and the stretched pulses are applied to the grids of a long-tailed pair. The difference voltage at one of the anodes is then used to control the reflector, the frequency response of the detector outputs and of the reflector voltage being shown in Fig. 9. In addition to the difference voltage available at this anode, a signal also appears across the common cathode load. This signal being proportional to the sum of the two inputs, the frequency response to this point is similar to that of a normal band-pass coupling. Thus a negative signal appears at this cathode whenever signals are received down the a.f.c. chain. A d.c. coupling connects this cathode to the grid of the phase-shift oscillator, so that the latter is biased-off when i.f. signals are received down the a.f.c. amplifier.

The sequence of operations on switching on the equipment is as follows: before the transmitter commences firing, the phase-shift oscillator will be sweeping the klystron frequency through its electronic-tuning range. When the transmitter begins to fire, and the oscillator frequency approaches the correct tuning point, a tuning-correction voltage will be received from the anodes of the long-tailed pair, while a negative voltage appears at the cathode. This latter voltage will bias-off the phase-shift oscillator while the reflector will be maintained at approximately the correct potential by the error voltage at the anode. The phase-shift oscillator, when biased-off, will tend to apply a voltage surge to the reflector, and if the klystron is to stay on frequency, the main a.f.c. loop must be able to cancel this surge. This can be achieved if the frequency of the phase-shift oscillator is sufficiently low, so that the anode will not rise too quickly to h.t. potential.

The second circuit avoids the complicated situation which arises when the Cockroft a.f.c. circuit approaches the correct frequency during the search. Examples of this circuit are to be found in a number of systems developed both in this country and in the U.S.A. (In the "Microwave Receivers" volume of the Radiation Laboratory series an arrangement of this type is repeatedly but quite inaccurately referred to as the 'diode phantastron' circuit.) In the usual form of this circuit, shown in Fig. 16, the reflector is coupled to the anode of a Miller integrator which forms part of the normal a.f.c. loop. The bias conditions are so adjusted that the anode of the integrator 'bottoms' if no signal is received from the discriminator and pulse-stretching circuits, and transitron feedback is provided between screen and suppressor of this valve. Thus, in the absence of a signal from the discriminator, this integrator will act as a free-running sawtooth generator. The transitron operating conditions are so chosen that the amplitude of the sawtooth is somewhat greater than that required to cover the electronic-tuning range of the klystron.

When the transmitter is switched on, a signal will be received at some point in the search cycle. When a negative output is received from the pulse-stretcher, the Miller run-down of the anode

will be slowed down and finally arrested at a suitable anode potential. The valve will thereafter function as an integrator in the normal manner. A residual tuning error must exist in order to hold the reflector at a suitable potential but, by making the steady-state gain sufficiently high, this error can be kept down to negligible proportions.

The transitron circuit is not highly satisfactory in a Service equipment, since it makes use of various valve characteristics which are not controlled. The difficulties are increased when it is attempted to maintain only a small amplitude of oscillation about a mean potential suitable for a particular klystron. It is, therefore, usual to provide two preset adjustments of the transitron operating conditions. It is, however, possible to replace this somewhat crude transitron oscillator by a more 'designable' circuit in which the limits of the anode excursion are more clearly defined.

In many equipments the frequency range over which the search is conducted is comparable with the i.f. It is then found that signals are received from the a.f.c. mixer, not only when the oscillator is approximately on tune, but also when the difference between the l.o. and transmitter frequency is any submultiple of the desired i.f. This effect is due to harmonics of the difference frequency generated in the mixer. The amplitude of these harmonics falls off as their order increases, so that it is unusual for trouble to be experienced from harmonics other than the second and third.

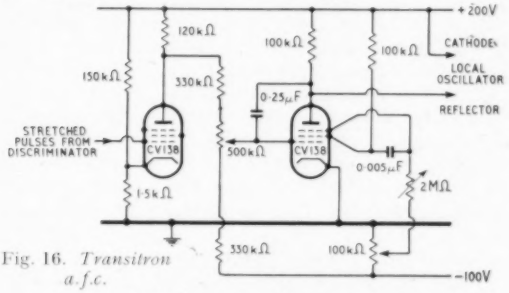


Fig. 16. *Transitron a.f.c.*

It is essential to arrange the search so that the a.f.c. locks only to the wanted signal, and not to the weaker harmonic responses. Even when the input levels are correctly chosen, the amplitude of these harmonics differs from the wanted signal by only 15–20 db (perhaps 25 db if a balanced a.f.c. mixer is used), so that if the relative amplitude of the signals is the only means of distinguishing between them, the margin of safety is quite small. For this reason it is not unusual to provide a preset adjustment of the signal level at which locking is to occur. Modern practice, however, is to make use of the improved tolerances on mixer crystals and better waveguide techniques to remove the necessity for this adjustment.

Reference has been made to the effect on the performance of the tolerances on the gain round the feedback loop of an extrapolating a.f.c. system. A list is given below of the more significant of the many factors contributing to variations in the gain. The importance of some of these factors depends on whether the l.o. or the transmitter drive to the a.f.c. mixer is the greater of the two, since, to a first approximation, the output will be directly proportional to the smaller of the two signals and independent of the larger.

TABLE 2

Typical tolerances on:	*Contribution to R*
(a) Transmitter power reaching a.f.c. mixer (b) Local-oscillator drive	1·8
(c) Gain of a.f.c. circuits	2
(d) Slope of electronic-tuning characteristic (at centre of mode)	1·5
(e) Variation in slope over mode	2·5

The tolerances on l.o. and transmitter drive are likely to be much the same but, other things being equal, the final value of R (ratio maximum/minimum possible loop gain) will be lower if the local-oscillator drive is the smaller of the two. If the l.o. is providing the signal controlling the mixer output, the fall in the l.o. output as the side of the mode is approached will tend to compensate for the increase in slope of the electronic tuning characteristic of the klystron.

If the transmitter drive is taken as the lesser of the two mixer inputs, a probable final value of R is 14 while if the l.o. is the smaller, the reduction in gain at the edges of the mode will reduce R by about 1·3; i.e., to 11. In a wideband system these figures can readily be exceeded because of the necessarily greater variation in the drive to the mixers and in the loading of the oscillator cavity, the Q of which determines the slope of the electronic-tuning characteristic.

It is possible to provide an a.g.c. system round the a.f.c. loop but this a.g.c. cannot take account of the slope of the electronic-tuning characteristic, and its application to the above system is not justifiable.

Wide-Range A.F.C.

The frequency range over which a reflector a.f.c. can hold in is limited by the electronic-tuning range of the klystron. In some higher-frequency equipments this range is barely adequate to deal with probable variations in the frequency of the magnetron and the klystron.

There is a limit to the electronic-tuning range that can be achieved before the efficiency of the

klystron falls too low and its noise output becomes too high, and it is, therefore, often desirable to adopt some technique which does not confine the operation of the a.f.c. system to the limits of the electronic-tuning range, especially as war-time experience showed the difficulty of producing a reliable electronic a.f.c. for airborne radars. There is, therefore, much to be said for applying the policy, already adopted in the waveguide system, of designing circuits to operate over the widest possible frequency range without adjustment.[15]

Various attempts have therefore been made at producing an a.f.c. system having the wider tuning range obtainable by mechanical means. A purely mechanical a.f.c. is not capable of dealing with scanner pulling, and similar short-term fluctuations in transmitter frequency, and most proposals involve some combination of electronic and mechanical tuning.

One arrangement consists essentially of a tuning motor which searches a wide band of frequencies until the wanted signal is located. The motor is then switched off, a reflector a.f.c. is allowed to take over, and is expected to deal with all subsequent variations. This system cannot be said to have overcome in any very satisfactory manner the limitations of a purely electronic a.f.c., and it is now more usual to use both forms

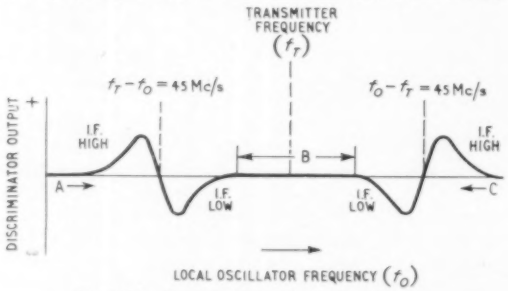


Fig. 17. *Output from i.f. discriminator.*

of correction simultaneously. If an attempt is made to connect two normal servos in parallel, both using the discriminator output to control the frequency, it will be found that the design problems become insuperable, since each servo receives an input only by virtue of the defects of the other, and it becomes impossible to determine either the steady-state position of the mechanical tuner or the condition that will ever be reached.

This problem can be avoided if some technique is adopted which divides the responsibility between the two servos in a less ambiguous manner. The first suggestion on these lines appears to have come from G. H. Nibbe, of Radiation Laboratories, M.I.T., who proposed to attempt to supplement the thermal tuning of a local oscillator with the electronic-tuning characteristic of the reflector.

Nibbe suggested that the design problem could be tackled by employing a reflector a.f.c. to hold the frequency at the desired value while the thermal tuner was controlled by a circuit which attempted to maintain the local-oscillator power output at its maximum value. Fig. 11 shows that if the mechanical tuning is varied while the frequency is kept constant by means of the reflector, the power output will fall off as the reflector voltage departs from the value corresponding to the centre of the electronic-tuning range. Thus the combined action of these two control systems will maintain the oscillator on tune with both the mean reflector potential and the mechanical tuning near their optimum positions. This technique has been successfully adopted for the control of a motor-tuned oscillator.[16]

In a wide-range a.f.c., many of the problems discussed earlier are accentuated and some new ones are introduced. For example, the tolerances on the local-oscillator output and on the slope of the electronic-tuning characteristic must both increase with the tuning range. This means a reduction in the response speed of the system, and the increased tolerances on the l.o. output may add to the difficulties with harmonics. In practice, however, it is probable that the theoretical loss in performance of a wide-range system is considerably less than the losses in performance of a narrow-band system due to errors in the adjustment of the preset controls. All that has happened, on changing over to a wide-range system, is that much of the responsibility for the performance of the equipment has been shifted back from the maintenance mechanic or equipment user to the designer.

Some new problems arise when designing a wide-range search system. First, there are, as shown in Fig. 17, two l.o. frequencies which give rise to a beat at the desired i.f., but the sense of the feedback voltage obtained from a conventional i.f. discriminator can only be correct for one of them, thus complicating the search problem. Another difficulty is that the process of searching the entire tuning range may take considerably longer than in a narrow-band system, times of the order of 5–10 secs being not unusual. This delay is not important on first switching on but, particularly in high-power systems, there may be momentary interruptions in the transmitter output due to arcing, etc. If the search oscillator begins to run when these interruptions occur, and the system takes 10 secs or so to locate the transmitter frequency after the transmitter output is restored, the effect of these disturbances on the performance of the equipment will be greatly exaggerated. It is therefore necessary to design

the control circuits so as to minimize the time wasted by such interruptions.

To tune a klystron over a wide range of frequencies, it is necessary to vary the physical dimensions of the cavity by some means. The first determined attempts at solving the wide-range a.f.c. problems were made in the U.S.A. and resulted in the development of a number of thermally-tuned local oscillators and some a.f.c. circuits for use with them. Improved techniques have now completely displaced thermally-tuned valves and the problems of thermal tuning will not be discussed further.

Reference has already been made to the difficulties of producing a purely mechanical servo-mechanism to provide the a.f.c. correction. In the present state of the art, it is not easy to see how these difficulties can be overcome to make possible a good a.f.c. performance. Work has therefore been devoted mainly to the possibility of using a mechanical tuner to conduct the initial search, and to handle long-term drifts, while a reflector a.f.c. system deals with scanner pulling, power supply 'flutter', microphony, and any other short-term fluctuations.

Two possible techniques have been described briefly above. The second of these uses the discriminator output to operate a normal reflector a.f.c. system (either 'instantaneous' or 'extrapolating'), while a separate circuit adjusts the mechanical tuning to keep the power output of the local oscillator (measured in terms of the direct current in a crystal) at or near its maximum value. This ensures that the mechanical tuning is set to approximately the correct mean frequency; i.e., that the mean operating point is in the centre of the 'reflector mode'.

Since the power output of the oscillator will vary with frequency it is desirable to maintain the crystal current at its maximum value rather than at some standard figure. This can be done by allowing the mechanical tuning to 'hunt' continuously about the optimum, a circuit being used to reverse the motor direction whenever the crystal current begins to fall perceptibly. In the absence of a signal from the discriminator, the motor is arranged to sweep to and fro over the tuning range until the desired frequency is located.

It will be appreciated that in normal operation the local-oscillator frequency is not varying to any significant extent during this motion of the mechanical tuner, since the reflector a.f.c. is simultaneously adjusting the reflector voltage so as to maintain the i.f. at the desired value. The change in crystal current to which the motor control reacts is, in fact, directly due to the change in reflector voltage rather than to the change in mechanical tuning.

Readers who are aware of the considerable complications which arise in circuits for use with the thermally-tuned valves will notice that there is little difficulty in searching with this reflector-cum-mechanical system, due to the separation of the two functions of searching (performed by the motor) and locking (performed by the reflector system). During the search period the crystal-current circuit is rendered inoperative, and the motor is made to run continuously through the tuning range. If the end of the range is reached the direction of travel is reversed, either by means of a limit switch or a cam. Fig. 18 shows the manner in which the frequency of the oscillator will vary as the motor searches over the band. When the i.f. differs widely from the correct value, no output will be received from the discriminator controlling the reflector voltage, and the oscillator frequency will be substantially the resonant frequency of the cavity. If the 'wrong' channel is reached first, the reflector voltage will begin to vary so as to oppose the change in frequency. When the limit of its control is reached, the frequency will jump discontinuously to the other side of the discriminator curve, when the cavity will again take complete control of the frequency. As the 'right' channel is reached, the frequency of the oscillator will be pulled-in to the value giving the desired i.f., and it will be held at this point over a range of settings of the mechanical tuner. The search is now at an end, and the motor control circuit may be allowed to take control of the motion of the tuning. This transition may be initiated by a valve which is biased-off when

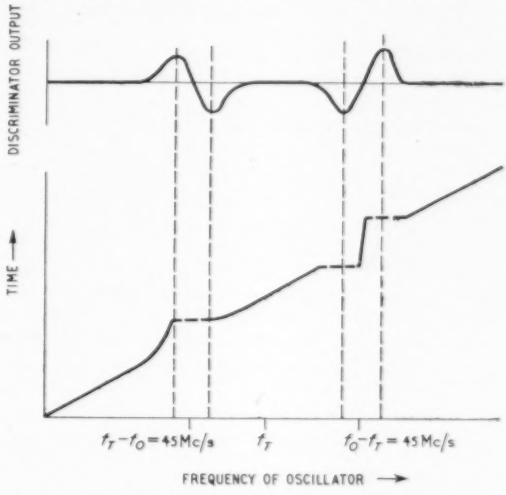


Fig. 18. *Variation of oscillator frequency during search.*

signals are received in both halves of the discriminator, since this means that the i.f. is in the cross-over region, and this only occurs when the wanted channel has been reached. Thus it is necessary to provide a circuit similar to that used to stop the phase-shift search oscillator in the Cockroft circuit of Fig. 15. It is necessary to provide sufficient loop gain to ensure that in the neighbourhood of the 'wrong' channel, the limit of the klystron's electronic-tuning range is reached before the frequency comes sufficiently near to the peaks of the discriminator to operate the search-stopping mechanism.

The absence of any wrong-channel problem makes possible the use of much simpler circuits than are required to obtain even an indifferent performance from a thermally-tuned valve. With a typical motor control circuit, and an 'extrapolating' a.f.c., seven valves are required after the discriminator. This is a saving of two valves over the Nibbe-Durand circuit[7] which gives a greatly inferior a.f.c. performance.

To maintain oscillations over a wide tuning range, it is necessary to gang a potentiometer to the tuning mechanism, in order to maintain the correct mean reflector voltage as the frequency is varied.

The crucial point in the design of the mechanical system and the motor control circuit is the choice of the speed at which the mechanical tuner is driven by the motor. There are four main factors:—

(a) The life of the mechanism will decrease with increasing motor speed.

(b) The control circuit is operated by the derivative of the crystal current. The faster the drive the lower the gain required in the control circuit, and therefore the less the danger of trouble from microphony, hum on the supply lines, etc.

(c) The rate of mechanical tuning in Mc/s per pulse should not be too high, otherwise the reflector a.f.c. may lag noticeably with a subsequent loss in tuning accuracy. This limitation is usually only serious at low p.r.fs.

The drive may be either continuous, or discontinuous. The advantage of a discontinuous drive is that a high rate of change of current is produced for a small amount of wear in the tuning mechanism, but there does not appear to exist a satisfactory motor which can produce a discontinuous motion in both directions. The discontinuous drive may, of course, be obtained by other means, for example, by switching the power supplied to the motor, or by using a Geneva mechanism. The former involves some circuit complications, and it is very doubtful whether a motor of normal design would not give a longer life when running continuously. The Geneva mechanism involves some mechanical complication, but there is plenty of evidence from their use in film projection, etc., that a design is possible which has a perfectly satisfactory life.

In the design of the motor control circuits, it is necessary to take account of the shape of the waveform which the crystal-current amplifier is to handle during normal operation of the circuit. For a discontinuous drive and a low drive/rest ratio the nature of the waveform is fairly obvious. The waveform is not so obvious when a continuous drive is used, since it is a function of the acceleration characteristics of the motor. A useful empirical approach is to construct the drive mechanism and to complete the reflector a.f.c. loop, but to arrange to reverse the motor by a key which is operated by hand whenever the crystal current is observed to drop to the limit within which it is intended that the circuit should maintain the current. The frequency with which the key is depressed should be measured and the amplifier is designed on the assumption that the input is a sine wave of this frequency. It is advisable to cut off the frequency response of the amplifier at 2 to 4 times this frequency, but the low-frequency response of the amplifier should be maintained down to at least one-tenth of the fundamental frequency, so that the only short time-constant is that feeding the trigger circuit at the end of the amplifier. This differentiating time-constant is probably best made roughly equal to a fifth of the time interval needed for a complete cycle of the motor operation.

Two other circuits[17,18] which have been tried under laboratory conditions use the discriminator output to control both the motor and the reflector a.f.c. The reflector a.f.c. system shown in Fig. 19 has a threshold tuning error, below which the error voltage is too small to overcome the bias on either of the diodes. Thus the reflector a.f.c. will reduce the tuning error only to this threshold value, which is not greater than about 10% of the receiver bandwidth, so that the presence of the residual error need not affect the performance of the equipment. The sign of the residual error is the same as that of the tuning error which the reflector a.f.c. is being called on to correct, so that if a tuning motor is controlled by a 'bang-bang' circuit actuated by the error voltage from the discriminator, this residual error can provide sufficient information to enable the motor to take over from the reflector a.f.c. as much of the task of keeping on tune as mechanical limitations permit.

In the second experimental system the reflector correction is applied through an a.c. coupling, so that, in the steady-state condition, the entire

correction is applied by the mechanical tuner, the reflector correction being reserved to deal with transients.

The transient response, for small errors, is identical with that of a normal extrapolating a.f.c. provided that the reflector coupling time-constant is long compared with the response time of the a.f.c. loop. Since this latter should not exceed about five pulses, whereas the coupling time-constant should be greater than the response time of the mechanical servo system, this condition is easily met.

A formal analysis of the stability conditions of a complete a.f.c. of this type, taking due account of the modification of the behaviour of the a.f.c. due to the finite p.r.f., offers no theoretical difficulty but is extremely laborious. It has already been shown, however, that the reflector a.f.c. will have a response time very much less than that of the mechanical system, so that the analysis may be simplified without much loss of accuracy, by discussing the response of the mechanical system on the assumption that the response of the reflector a.f.c. is instantaneous. The virtual impossibility of allowing for the effect of 'stiction' and backlash sets a limit to value of these calculations.

These last two circuits have the advantage that they can be used with a local oscillator having an irregular or discontinuous 'mode'. A disadvantage is that the reflector-voltage tracking circuit must work very well if the local-oscillator drive is not to vary considerably over the mechanical-tuning range due to variation of the position of the operating point on the reflector 'mode'. The 'crystal-current' circuit will always settle down with the operating point in the centre of the mode; the reflector-voltage tracking need only be good enough to maintain some sort of oscillation during the search. Further research is also necessary to determine whether the 'bang-bang' circuit used with the second scheme is capable of maintaining the tuner in the correct position in the presence of appreciable scanner pulling, which will greatly confuse the information reaching the trigger circuit.

Present experience, admittedly somewhat limited, suggests that the 'crystal-current' scheme is capable of giving perfectly satisfactory results with any oscillator having a reasonable 'mode'. Freedom from discontinuities over the electronic-tuning range is also necessary for the satisfactory functioning of the reflector a.f.c. This limitation is therefore inherent in any a.f.c., and the use of this type of system does little more than emphasize it. It is believed that this circuit will offer a satisfactory solution to almost any wide-range a.f.c. problem. The other two schemes have not been developed to a state where they can be recommended with any confidence, but they are mentioned here for the sake of completeness.

Some Practical A.F.C. Circuits

Previous writers[12] have commented on the remarkable diversity of circuits which have been used for radar a.f.c. No attempt will be made here to catalogue these circuits or discuss their genesis. A few outlines will be given of some recent designs which avoid the more serious defects of earlier circuits, illustrate minor design problems, and which may help readers to judge the amount of complexity involved in the techniques under discussion.

Fig. 19 shows a simple electronic a.f.c. with a 'transitron' type search. In the absence of a signal from the discriminator, V_4 and V_5, which replace the conventional transitron, will apply a saw-tooth sweep to the reflector. The detailed mechanism is as follows: V_4 is so biased that, in the absence of an input signal, its anode will run down till V_5 begins to conduct. The screen current of V_4 will then begin to rise, and the feedback to the grid of V_4 through the blocking-oscillator transformer T_2 and V_5 will cause V_4 to swing into heavy screen and grid current. The screen of V_4 will fall nearly to earth potential, and the cathode of V_5 will

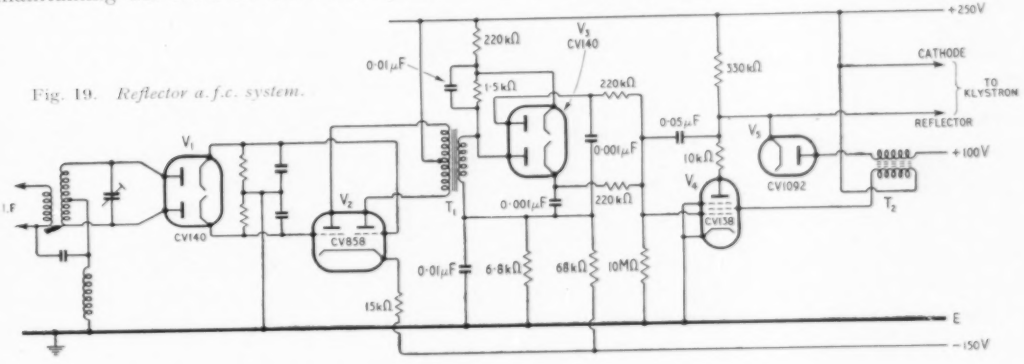


Fig. 19. *Reflector a.f.c. system.*

therefore rise by an amount nearly equal to the h.t. voltage multiplied by the turns ratio of T_2. This is chosen to suit the sweep voltage required by the klystron. Soon after the peak of the swing has been reached, V_5 will cut off and the heavy negative-feedback through the Miller capacitor will take over. V_4 and V_5 thus generate a saw-tooth sweep for the reflector.

When signals are received, the discriminator diodes V_1 will charge their load capacitors during the transmitter pulses. The resulting video signals, somewhat lengthened by the discharge time-constant of the diode loads, are applied to the grids of a long-tailed pair, V_2. The current pulses in the two anodes are subtracted from each other in the pulse transformer T_1, which has a 2 : 1 step down from each anode so that a peak current of about 20 mA is available from the secondary. The voltage gain of V_2 is very low, because of the low anode-load impedance, so that the use of triodes in this stage presents no difficulty.

To complete the process of converting the error pulses to d.c. V_3 is provided. The twin diodes are connected back-to-back with a volt or so of bias on each. In the absence of an error pulse of amplitude exceeding this bias, both diodes remain non-conducting, and the anode of V_4 can only drift very slowly. In the presence of an appreciable tuning error, one section of V_3 will conduct, charging its load capacitor to the peak output voltage of T_2. This voltage will be integrated by V_4, producing a suitable change in the reflector voltage of the klystron and hence a compensating shift in its frequency. The discharge time constants of the load capacitances and resistances of V_3 should not exceed the time interval between two successive pulses, so that the complete loop will behave substantially as the 'single time-lag' system discussed in the Appendix, which also explains the factors influencing the choice of the speed of response of the feedback loop to a tuning error. In practice this speed, when chosen, is adjusted by the choice of the Miller feedback capacitance round V_4.

This arrangement differs in a number of important particulars from many published circuits of this type. First, there are no inter-stage d.c. couplings, so that it is not necessary to provide presets to take up component tolerances. In addition, the blocking-oscillator type of feed-back avoids the drawbacks of the transitron, and the symmetry of the video portion of the discriminator eliminates many of the troubles which otherwise arise from the capacitance unbalance in the diode loads. The tolerances on the reflector characteristics of the klystron, a CV2116, are such that no adjustment of reflector voltage is needed even when the klystron is replaced.

Fig. 20 shows the circuit used by C. Baron in his pioneer work on the 'instantaneous' a.f.c. Three i.f. stages V_1, V_2 and V_3 feed a Foster-Seeley

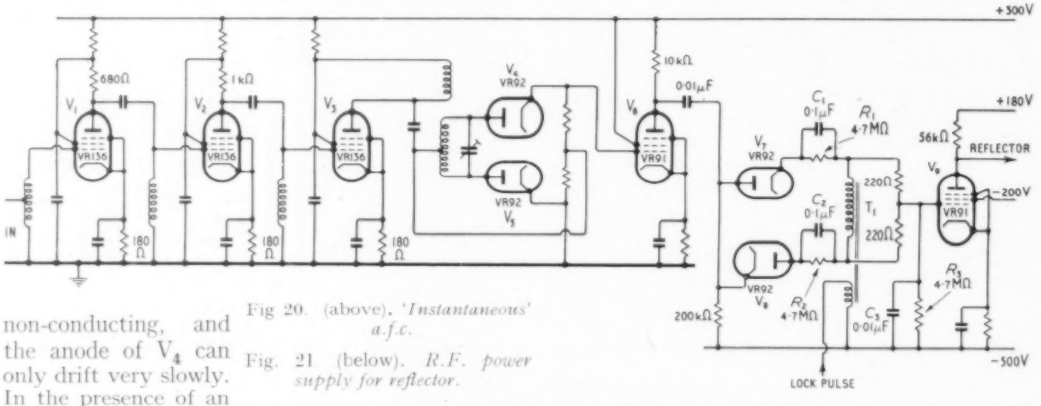
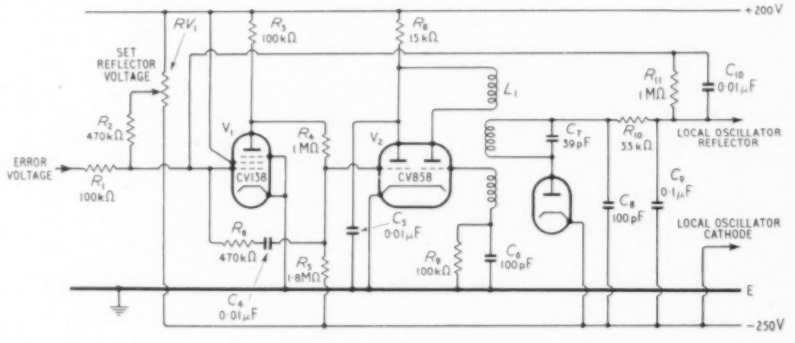


Fig. 20. (above). *'Instantaneous' a.f.c.*

Fig. 21 (below). *R.F. power supply for reflector.*



discriminator having a peak separation of 22 Mc/s. The discriminator output is amplified by the video stage V_6 and fed into the 'memory' capacitor C_3 through a 'clamp' circuit consisting of a pair of diodes V_7 and V_8 which are switched on when a 'lock' pulse is applied to the transformer. The current taken by the two diodes produces a more or less steady bias across the capacitors C_1 and C_2, so that the diodes will be non-conducting in the intervals between locking pulses. Thus C_3 will remain charged when the locking pulse ends. (The time constants C_1R_1, C_2R_2 and C_3R_3 are very long compared with the reciprocal of the p.r.f.)

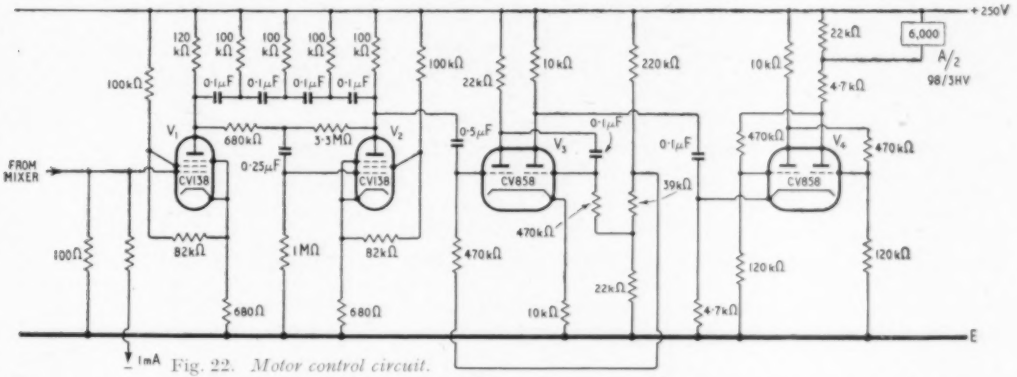


Fig. 22. *Motor control circuit.*

After further amplification the voltage across C_3 is applied to the reflector. A mathematical analysis of the stability conditions for this circuit is given in the Appendix. It is not proposed to enter into a lengthy discussion of the circuit problems peculiar to the 'instantaneous' a.f.c. system, such as that of devising a suitable clamp circuit. The Appendix shows that, except possibly for very short pulses, the design does not present any major difficulty. It is also shown that this system is less sensitive to circuit tolerances than the extrapolating type of a.f.c., since the instantaneous scheme can always be designed to cope with circuit tolerances, provided that the a.f.c. circuit bandwidth is increased sufficiently.

The circuit of Fig. 20 is open to a number of minor criticisms, but it is included here because examples of this type of circuit are rare, and because this example is of some historic interest.

A recurring problem in the design of reflector a.f.c. systems is set by the necessity of providing a d.c. coupling between the last a.f.c. stage and the reflector of the klystron. In most early equipments, as in the examples given above, the klystron was operated with its cavity at a positive potential with respect to earth. The cathode could then be directly connected to the anode of a control valve. This arrangement usually complicates the power supply problems, and has the disadvantage that the metal body of the klystron and the mechanical tuner are at a dangerous potential. There are also insulation problems, in the coupling between the klystron output and the mixers, for example. These objections become even more cogent if a high-voltage klystron is used.

These difficulties can be avoided if an r.f. power supply is used for the reflector. A suitable circuit is shown in Fig. 21. One section of a double triode (V_2) maintains oscillation at a frequency of about 500 kc/s in L_1. The voltage across a secondary winding is rectified by the diode and applied between cathode and reflector of the klystron. This voltage can be varied by means of the grid of the other triode section of V_2, which modulates the oscillator by varying the anode voltage. The grid of V_2 is d.c. coupled to the amplifier stage V_1. The error voltage, consisting of stretched pulses from the discriminator, is fed into the 'see-saw' input circuit of V_1, the other arm being taken to the reflector feed. Thus the steady-state gain from V_1 to the reflector is fixed by the ratio of R_{11} to R_1. V_1 is necessary mainly to provide a phase-reversal to allow negative-feedback to be applied round the loop. The rate of response of the reflector voltage to an error is controlled by the integrating capacitor C_{10}. The risk of phase-shift oscillations round this local-feedback loop is avoided by making the filter time-constant (C_9R_{10}) long compared with the other time-lags in the system (e.g., R_8C_8 and R_9C_6), and by restricting the a.c. loop gain by further feedback (R_6C_4) round V_1. To maintain the correct d.c. conditions at the grid of V_1, it is necessary to feed in some d.c. from the positive h.t. line. If klystron tolerances make it necessary, an adjustment (RV_1) can be provided to set the mean reflector voltage to suit the individual klystron.

By virtue of the heavy feedback round the

loop, this arrangement is quite stable and linear, and the reflector voltage can be controlled over a much wider range than is ever necessary. Although this technique requires an extra valve in the a.f.c. circuit, the simplification of the power supplies usually proves a more than adequate recompense.

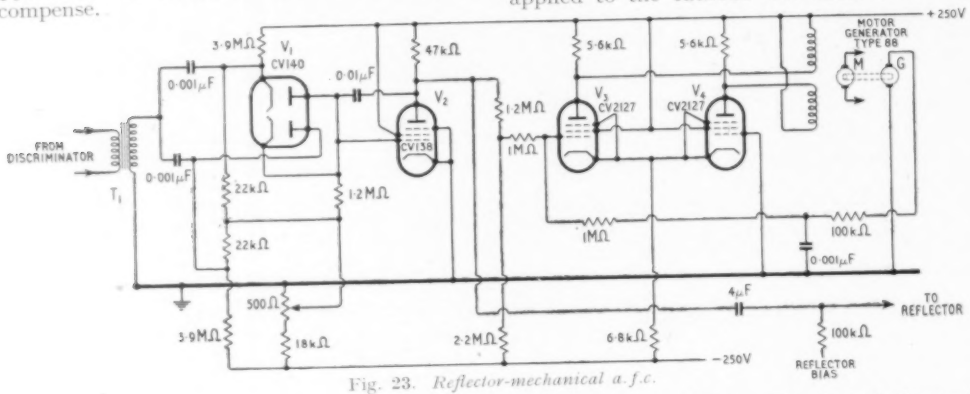


Fig. 23. *Reflector-mechanical a.f.c.*

The circuit illustrated in Fig. 21 was designed for use with a low-voltage klystron, but the technique can be even more profitably applied to high-voltage valves. If it is not convenient to take feedback directly from the reflector to the grid of V_1, it is possible to provide duplicate output windings and rectifiers, one of which develops a voltage between the cathode and reflector of the klystron, while the output of the other is earthed at one end and is used to provide the feedback. Such an arrangement overcomes the main objection to the use of high-voltage oscillators in conjunction with a.f.c. systems.

Fig. 22 shows an experimental circuit used for controlling the tuning motor of a reflector-cum-mechanical a.f.c. using the crystal-current scheme. The direct current in the a.f.c. mixer (obtained from the input to the a.f.c. i.f. amplifier, after filtering, develops a voltage across a 100-ohm resistor in the grid of an amplifier stage V_1. The resultant voltage, after further amplification by V_2, is differentiated and applied to the grid of a trigger circuit V_3, which trips whenever the mixer current falls sufficiently. To prevent this circuit being tripped by hum, microphony, or video pulses from the a.f.c. mixer, the amplifier stage V_2 is arranged to have a frequency response which commences to fall above about 10 c/s. This fall is brought about by the phase-shifting feedback network between the anodes of V_1 and V_2. The trigger circuit V_3, when tripped by a signal from V_2, produces a pulse of about 0·05 sec duration which is differentiated and fed into the cathode of the 'divide-by-two' circuit based on V_4. This is a d.c. coupled multivibrator with a relay in one of the anode circuits. Briefly, the mechanism of division is as follows:—

Negative pulses applied to the cathode of V_4 produce little effect since whichever section of V_4 happens to be conducting will merely be driven a little further into grid current; positive pulses applied to the cathode will cause a rise in the anode potential of the conducting triode, applying a positive swing to the grid of the non-conducting section. This grid voltage will be much greater than that at the cathode because of the gain of the conducting triode; the net result being to turn-off the formerly conducting section and vice versa, further reversal during the pulse being prevented by the inductive surge developed across the relay winding at the first reversal.

When V_4, and therefore the relay A/2, changes its state, the direction of running of the mechanical tuner motor is reversed. (The motor wiring is omitted from Fig. 22.) The motor will thus 'hunt' to-and-fro about the position where the crystal current is a maximum. The local oscillator used with the circuit shown was an X-band valve, the 723A/B. The tuner was driven by a two-way 'stepper' motor running at a speed of about 2 or 3 steps per second, each step producing a frequency shift of about 2 Mc/s. The 723A/B is not mechanically very suitable for use in this way, but was used merely to try out the principles. The search mechanism consists of a relay in the reflector control circuit which operates when the reflector loop pulls in. When this relay is operated, the above circuit is allowed to control the motor, but when the relay is open, the tuner will be driven to-and-fro over its range under the control of a toggle mechanism which reverses the motor when the limits are reached. Normally, when the correct frequency is reached during the search, the a.f.c. will lock in, the relay will operate and the motor control circuit can take over.

Fig. 23 shows an experimental version of a wide-

range a.f.c. using an a.c. coupling to the reflector of the klystron. The latter was a high-voltage X-band valve, the CV129, whose tuning mechanism was driven by a Velodyne through a suitable gearbox. The gear ratio was chosen so that, at full speed, the tuner was driven at about 10 Mc/s per second. This is more than adequate to deal with any thermal drift. The CV129 has an electronic tuning range of only about 3 Mc/s, since this valve was not intended for use with an a.f.c., but the assembly of CV129, gearbox and Velodyne happened to be available in the laboratory.

An a.f.c. mixer, i.f. amplifier, and discriminator of conventional design, not shown in Fig. 23, supply error pulses to the transformer T_1. These error pulses pass through the diodes V_1, and are added to the output of V_2. The bias on the diodes prevents them conducting during the intervals between the pulses, but otherwise has little influence on the operation of the circuit. The output of V_2 is a.c. coupled to the reflector of the CV129, and also controls the input to the long-tailed pair (V_3, V_4) and hence the field current of the motor.

Before connecting the a.c. coupling to the reflector of the CV129, the mechanical a.f.c. was first tested alone. It was found that there was a marked tendency to instability, largely, it was believed, because of the considerable 'stiction' on the tuner shaft of the CV129, which provided the mechanical load on the Velodyne. This 'stiction' is abnormally high because the tuning mechanism of this valve is far from ideal for use with a motor drive, but high stiction is likely to be a feature of almost any such mechanism, because of the infinitesimal movement required.

On connecting up the reflector feedback, this was found to exert a marked stabilizing action on the system, which then behaved in a perfectly satisfactory manner. It appears quite likely that a simpler type of servo motor could have been used. The search system is again omitted from the diagram, but that described in the previous example is equally applicable here.

(To be concluded)

References will appear at the end of the last instalment.

NEW BOOKS

The Electromagnetic Field and its Engineering Aspects

By G. W. Carter, M.A., M.I.E.E. Pp. 360 + xvi. Longmans, Green & Co., 6 and 7 Clifford Street, London, W.1. Price 35s.

The author is Professor of Electrical Engineering in the University of Leeds, and the book is dedicated to his father, the late F. W. Carter, M.A., D.Sc., F.R.S., for many years Consulting Engineer to the B.T-H. Co. The object of the book is stated to be to give a thorough-going non-mathematical treatment of the theory of electromagnetism, a secure grasp of which has been made more than ever necessary by recent developments in engineering. Its first aim is to explain how electromagnetic apparatus behaves. One must not be misled by the term non-mathematical, for, as the author confesses, the treatment goes deeper than is usual in non-mathematical books, and the final chapter is devoted to 'the Principle of Relativity'. The rationalized m.k.s. system is used throughout. A very full list of symbols is given at the beginning and one receives a shock to note that, contrary to B.S.I. recommendations and general usage, the symbol for e.m.f. is $\mathcal{E}$ and that for electric force is E; this is not a misprint but is used throughout the book.

Chapter 1 consists of a brief historical introduction and a note on units. Chapter 2 deals with the electric field of charges in free space, and Gauss' theorem. In the formula $D = \epsilon E$, ϵ is called the 'electric constant' and when applied to free space it is called the 'primary electric constant'. Chapter 3 is entitled 'Conductors, Insulators and Capacitance', and concludes with a section on electrostatic screening. Chapter 4 deals with energy and the mechanical forces in the electric field, concluding with the electrostatic voltmeter and quadrant electrometer. In Chapter 5 we leave electrostatics and turn to electric-circuit theory and Ohm's law; the electrolytic tank method of plotting electric fields is described. In Chapter 6 we turn to magnetism, current loops and magnetic dipoles, magnetic potential and circuits, and Biot-Savart's law. What is generally called the absolute permeability of free space is called the primary magnetic constant and represented by the symbol η_0. The symbol μ is used for the ratio η/η_0 just as $\kappa = \epsilon/\epsilon_0$ is used for permittivity. Chapter 7 deals with the magnetic effects of iron. Different types of iron and steel are discussed and also various alloys and powdered cores; although ferrites do not appear to be mentioned, references are given to works on the subject, and it is explained that the account given is very incomplete. Chapter 8 is entitled 'electromagnetic induction', and deals with flux threading, flux cutting, slotted d.c. armatures, self and mutual inductance, etc. The energy and mechanical forces in the magnetic field are discussed in Chapter 9; the theory of the flux meter is given; hysteresis, eddy currents and Maxwell stresses are discussed. Chapter 10 is devoted to vector potential and its application to the calculation of self and mutual inductance. Chapter 11 deals with skin effect in a very thorough manner. Chapters 12 and 13 are devoted to Maxwell's equations, electromagnetic waves in space and in coaxial cables, waveguides, oscillating dipoles, the Poynting and Slepian vectors; one section is entitled 'the transition from low to high-frequency engineering'. The final chapter gives a brief discussion of the principle of relativity and its effect in electrical engineering.

A number of mainly numerical examples is given at the end of each chapter and the solutions are given at the end of the book.

In some cases the methods adopted by the author seem to be unnecessarily complicated, due to "the author's strong preference for the direct application of Neumann's law $\mathcal{E} = -d\Phi/dt$", which leads him to apply it where its application necessitates the picturing of fictitious circuits. A striking example of this is the

Faraday disc rotating in a uniform field parallel to the axis, so that every free electron experiences a radial force; this e.m.f. causes a p.d. between the axis and the periphery, and it seems quite unnecessary to explain it as being due to increasing flux linkage in a fictitious circuit. The author returns to it again in the chapter on relativity, and we are told that "the relativistic effect arises from the fact that the disc material is moving across the circuit line PQ". An interesting variant of the ordinary Faraday disc would be obtained by using for the external circuit a cylindrical can a little larger than the disc, with brushes all round the rim rubbing on the edge of the disc, and with the spindle prolonged to make contact with the centre of the bottom of the can. This would provide some further exercise for the Neumann law enthusiasts.

These remarks are not intended as an adverse criticism of the book, but rather the reverse, for they show that it is thought-provoking and attempts "to tie up a lot of loose ends which are commonly left untied". The aim of the book is stated to be "to give the reader a comprehensive view of the theory of electromagnetism, with constant reference to its engineering applications", and this it certainly does. It is a book that can be recommended to every serious student of electrical engineering.

G. W. O. H.

Transistoren

By Dr. M. J. O. Strutt. Pp. 166 with 121 illustrations. S. Hirzel Verlag, Zürich. Price 21 Swiss francs.

The author is Professor and Director of the Institute for Advanced Electrotechnics at Zürich. This volume is No. 18 of a series of monographs in German on electric communications published under the direction of Professor Feldtkeller, of Stuttgart. Work on transistors has been going on at Zürich since 1948, and this book is based on a course of lectures on the subject given during the session 1952–53. It takes into account, although perhaps not completely, the published literature on the subject up to the middle of 1953; references are given to 148 published papers. In the preface the author says that it is now certain that in wide fields the valve will be replaced by the transistor, as its advantages in the saving of power and space are obvious. The difficulties of standard production are, especially in the U.S.A., being overcome.

The book is divided into ten chapters, of which the first, entitled Introduction, is divided into seven sections with the following headings: (1) What is a transistor? (2) Characteristics of the point transistor, (3) Compared with those of a passive 4-pole, (4) Equivalent diagram of a point transistor, (5) Power gain, (6) Ways of connecting it as a 4-pole, (7) Stability. This opening chapter occupies 19 pages, and deals with all these points in a very thorough manner. There is an unfortunate muddle on pages 4 and 5 in the numbering of the equations; in five places the references to Section 2 should be to Section 3, but this is a minor detail and it is soon spotted.

The second chapter deals with the electronics of the material, slow moving electrons, electrons in atoms, the wave nature of the electron and then electrons in solid material, with special reference to semi-conductors. The third chapter is entitled Contact Electronics and deals with contact between different metals, between metals and semi-conductors, and between different semi-conductors, and concludes with a section on rectification at the contact. Having devoted three chapters to the electronics of the materials, Chapter IV deals with their use in point transistors, the current multiplication at the collector, the formation of the material at the surface, time of transit, capacitance, inductance and noise. Chapter V deals with surface or junction transistors of various types and phototransistors. Duality and Analogues, to which we devoted the Editorial in March 1952, form the subject of Chapter VI. Two chapters are devoted to the use of transistors in the early stages and in the late stages of amplifiers, the noise-to-signal ratio, bandwidth, distortion and comparison with valves. Chapter IX deals with the production of oscillations with back coupling and negative resistance, and the final chapter with methods of measuring the important properties of semi-conductors and of transistors. Although most of the material has already been published in some form or other, some chapters contain hitherto unpublished developments of this earlier work.

The book contains a very clear exposition and a very thorough discussion of almost all that is known or, at least, was known a year ago, of the properties and applications of this minute but revolutionary element of electronic equipment, but some readers may find it rather compressed; the Principles of Transistor Circuits, by R. F. Shea, recently published, contains about three and a half times as many pages, and yet the reviewer in the May *Wireless World* thought that some of the chapters were "disappointingly scanty".

G. W. O. H.

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(Communication from the National Physical Laboratory)

Values for April 1954

Date 1954 April	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	− 1·1	+ 2	NM
2	− 1·1	+ 2	+ 53·2
3	− 1·1	+ 1	NM
4	− 1·1	+ 1	NM
5	NM	NM	+ 47·5
6	− 1·2	+ 2	+ 45·2
7	− 1·2	+ 2	+ 42·9
8	− 1·2	+ 2	+ 42·5
9	− 1·1	+ 2	+ 38·6
10	− 1·1	+ 2	NM
11	− 1·1	+ 1	NM
12	− 1·1	+ 2	+ 34·3
13	− 1·1	+ 2	+ 33·4
14	− 1·1	+ 2	+ 31·8
15	− 1·3	+ 1	+ 30·6
16	− 1·3	+ 1	NM
17	− 1·3	+ 2	NM
18	− 1·2	+ 1	NM
19	− 1·2	+ 2	NM
20	− 1·2	+ 2	+ 23·6
21	− 1·2	+ 2	+ 22·1
22	− 1·2	+ 3	+ 20·4
23	− 1·2	+ 3	+ 19·3
24	− 1·3	+ 3	NM
25	− 1·3	+ 3	NM
26	− 1·3	+ 1	+ 15·3
27	− 1·2	+ 4	+ 14·2
28	− 1·2	+ 3	+ 13·3
29	− 1·1	+ 4	+ 12·0
30	− 1·2	+ 4	+ 10·3

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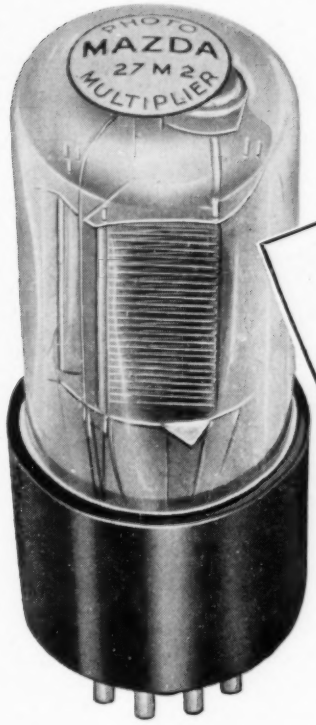

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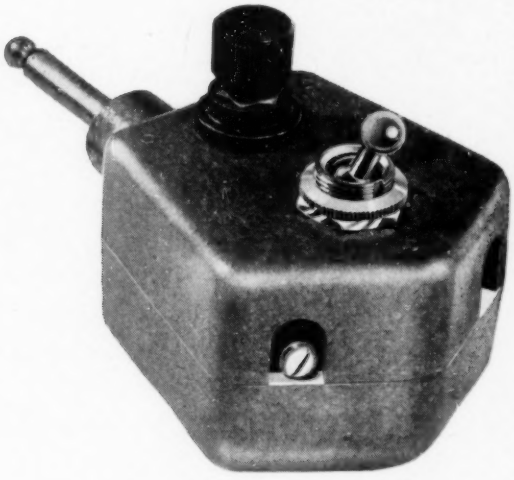

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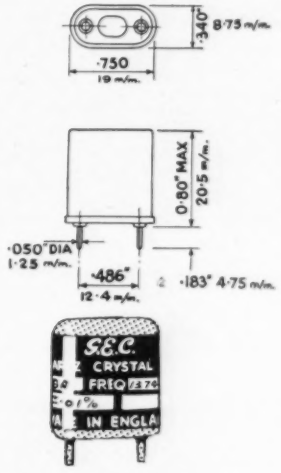

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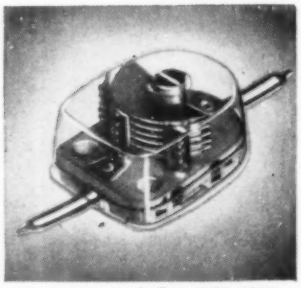

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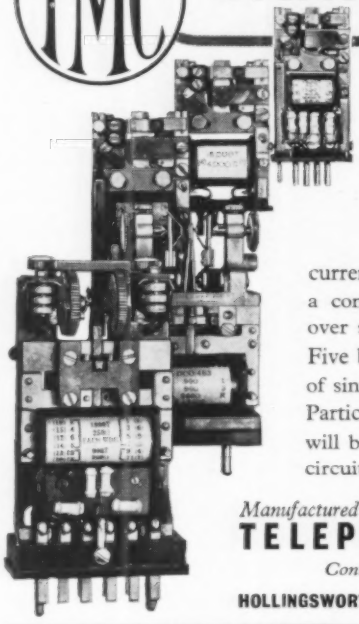

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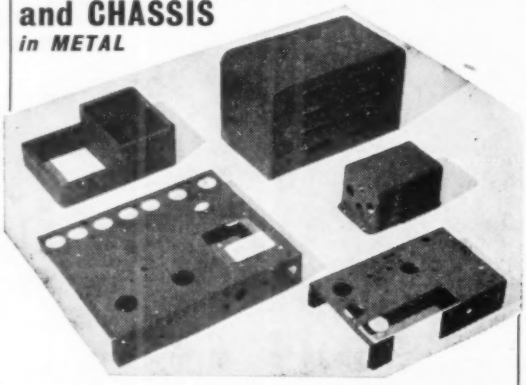

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Peak Inverse Voltage (V)	85	30	30	100	70
Minimum Reverse Resistance (kilo-ohms)	500 (at 70V)	100 (at 20V)	100 (at 20V)	500 (at 50V)	50 (at 50V)
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(Registered Office: Connaught House, Aldwych, London, W.C.2.)

Radio Division

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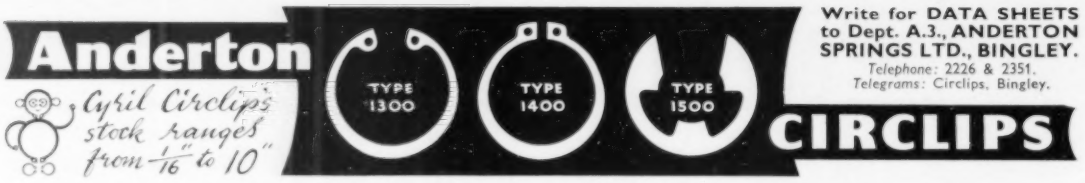

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The Patentee of British Patent No. 649,403 relating to "Apparatus for Detecting Faults in Electronic Circuits" desires to dispose of his rights either by way of sale or licence on a royalty basis. In the first instance please communicate with Eric Potter and Clarkson, Chartered Patent Agents, 317 High Holborn, London, W.C.1.

APPOINTMENTS—The engagement of persons answering these advertisements must be made through the local office of the Ministry of Labour and National Service, etc., if the applicant is a man aged 18-64 or a woman 18-59 inclusive, unless he or she or the employer is [excepted from the provisions of The Notification of Vacancies Order, 1952.

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(*a*) **One Engineer** required for work in connection with the development of electrical and electronic navigational aids. This includes the development of optical systems for fog detection apparatus, development of fog signals, audio monitoring equipment and automatic control and indication equipment.
Salary scale: £1,030 rising to £1,230 per annum.

(*b*) **Two Experimental Officers** required for work as set out in (*a*).
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(*c*) **Two Engineering Assistants** required for planning and progressing electrical installations, including engine—generating plant up to 40 KVA wiring and distribution and the design of small component parts.
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(*d*) **Two Laboratory Assistants** required for general laboratory work in connection with (*c*) above and radio/radar equipment.
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(*e*) **Four Radio Maintenance Assistants** required to maintain radio and radar equipment in shore stations and ships.
Salary scale: £407 rising to £550 per annum. (Minimum linked to age 25 years.)

Minimum Qualifications Required:
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For (*c*) Higher National Certificate in Electrical Engineering.
For (*d*) Higher School Certificate (Science) or equivalent.
For (*e*) a knowledge of the fundamental principles of radio and radar and practical experience in maintenance of use of such equipment.
All candidates must be medically fit and of British Nationality.

Appointments
A proportion of those appointed may be placed on the permanent established staff after a satisfactory probationary period.

Applications
Applications should be made in writing to the Secretary, Trinity House, London, E.C.3, not later than 18th June, 1954, stating age, occupation, qualifications and experience and enclosing copies of recent testimonials.

Electronic Engineer for Research Laboratory. Senior position for engineer with experience of research and design of special circuits at medium to low frequencies. Knowledge of Physics an advantage. Qualifications should include a good degree and preferably five years experience in similar post. Interesting and rewarding work with excellent prospects for advancement. (Ref. 73.)

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Write in detail, quoting Reference No. of position sought to: The Personnel Dept. (Technical Employment), De Havilland Propellers, Ltd., Hatfield, Herts.

The General Electric Co., Ltd., have vacancies at Stanmore for **Electronic** and **Electrical Engineers** for work in Development Laboratories. There are a few senior positions available to experienced men with a University Degree or equivalent qualifications and more junior positions for men of O.N.C. standard and/or with previous experience in development work. The work is interesting and there are openings in each of the following fields:—

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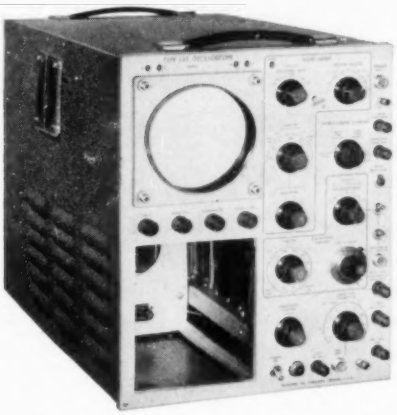

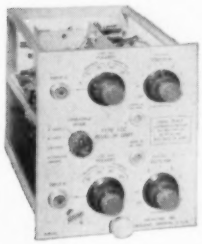

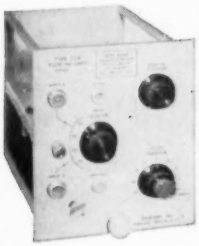

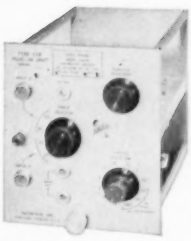

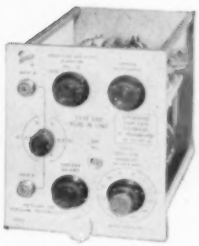

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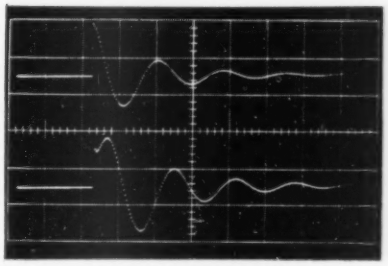

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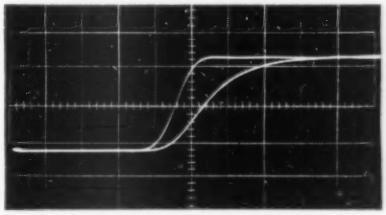
Response of two networks excited by a single pulse shows free-running operation of the Dual-Trace Unit in a one-shot application. A single 200-μsec/cm sweep is used for this display.



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Housing assistance available to successful applicants from the Greater London Area. Write, stating age, qualifications, experience, present salary and Ref. S.A.38A to The Personnel Manager, The English Electric Co., Ltd., The Airport, Luton, Beds.

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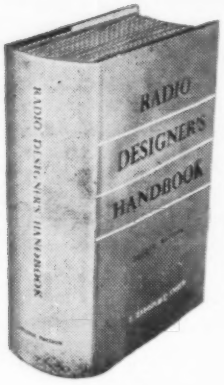

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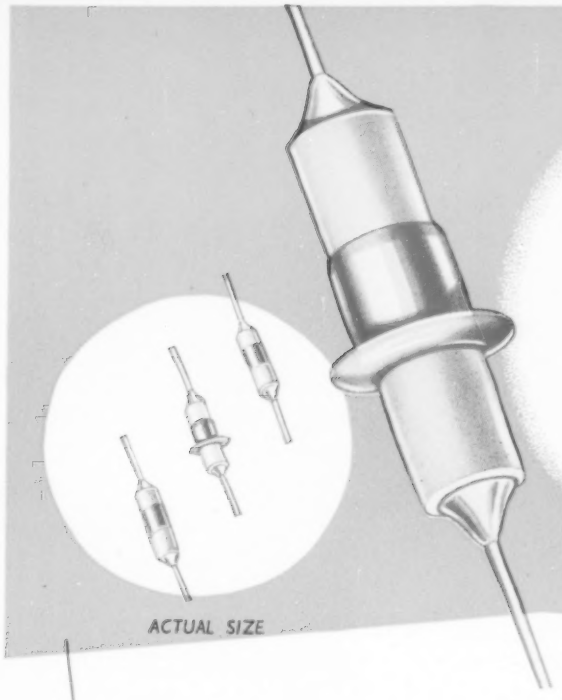

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